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ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

HIGH-TENSION LABORATORY
SUBMINIATURE-TUBE RELIABILITY TESTING
ELECTROMAGNETIC DELAY LINES
PRECISION-APPROACH AIRFIELD RADAR
FUNCTIONAL DIAGRAMS FOR ELECTRONICS IN TELEGRAPHY
CUBIC DISTORTIONS IN RING MODULATORS
MICROSTRIP FOR BAND-PASS MICROWAVE FILTERS
COAXIAL LINE WITH HELICAL INNER CONDUCTOR
UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM

Volume 32 **MARCH, 1955** **Number 1**

ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
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Volume 32 MARCH, 1955 Number 1

CONTENTS

High-Tension Laboratory ... 3
By Amund Braaten

Subminiature-Tube Reliability Testing ... 11
By Sterling Bickel

Electromagnetic Delay Lines ... 19
By H. G. Nordlin

Precision-Approach Airfield Radar ... 24

Functional-Diagram Approach to Electronics in Telegraphy ... 26
By E. P. G. Wright and D. S. Ridler

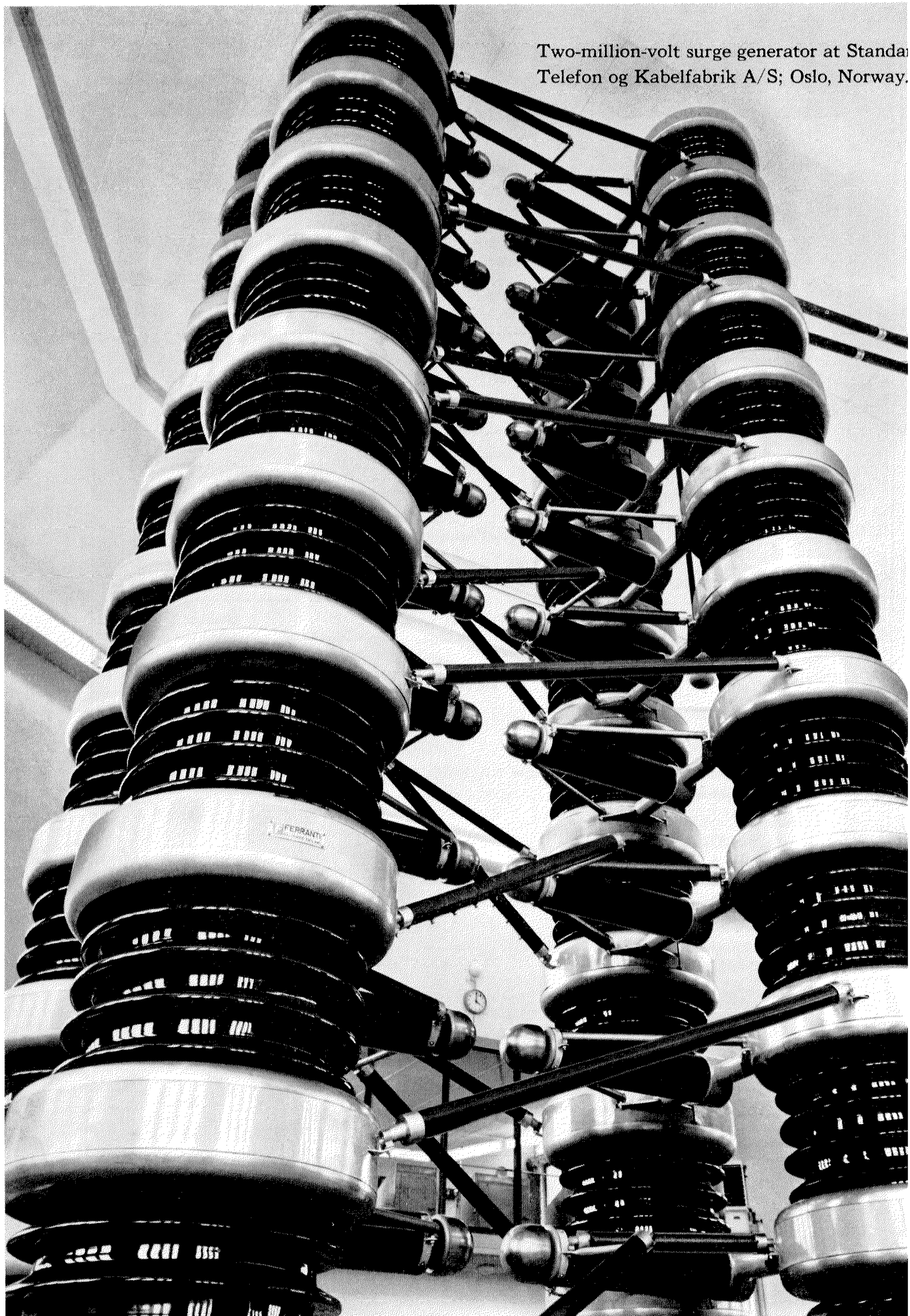
Cubic Distortions in Ring Modulators ... 43
By Leopold Christiansen

Microstrip Applied to Band-Pass Microwave Filters ... 52
By Maurice Arditi and Jack Elefant

Coaxial Line with Helical Inner Conductor ... 62
By William Sichak

United States Patents Issued to International Telephone and Telegraph System; August–October, 1954 ... 68

Contributors to This Issue ... 70



Two-million-volt surge generator at Standa
Telefon og Kabelfabrik A/S; Oslo, Norway.

High-Tension Laboratory*

By AMUND BRAATEN

Standard Telefon og Kabelfabrik A/S; Oslo, Norway

DEMAND for supertension power-transmission cables in connection with the development of large water-power resources in Norway after the second world war created the necessity for a modern high-tension laboratory for cable testing and development. For economic and defensive reasons, power plants are preferably built underground, which necessitates the use of high-tension cables for linking the transformers with the outside switching plant. Furthermore, there is a tendency to use cables instead of high-tension overhead lines in suburbs and built-up areas. There is also an increasing demand for high-voltage submarine cables for the electrification of coastal districts.

A simplified plan of the building is given in Figure 1, while Figure 2 shows a general view of the laboratory proper. The main building housing the high-tension equipment is 12-sided with a diameter of 98 feet (30 meters) and a floor-to-ceiling height of 46 feet (14 meters). The attached building has a floor space of 4300 square feet (400 square meters) and accommodates offices, physical laboratories for investigating insulating materials, instrument-maker's shop, and the control room for operation of the high-tension equipment.

The electrical installation consists of a 600-kilovolt alternating-current plant, a 2100-kilovolt surge generator, and a 1000-kilovolt direct-current set.

The alternating-current high-tension equipment is shown in Figure 3. It is capable of producing potentials up to 600 root-mean-square kilovolts and has a load capacity of 2400 kilovolt–amperes for 30 minutes. It consists of two units that may be connected in either series or parallel. Space is reserved for the future addition of a third unit that will raise the voltage to 900 kilovolts.

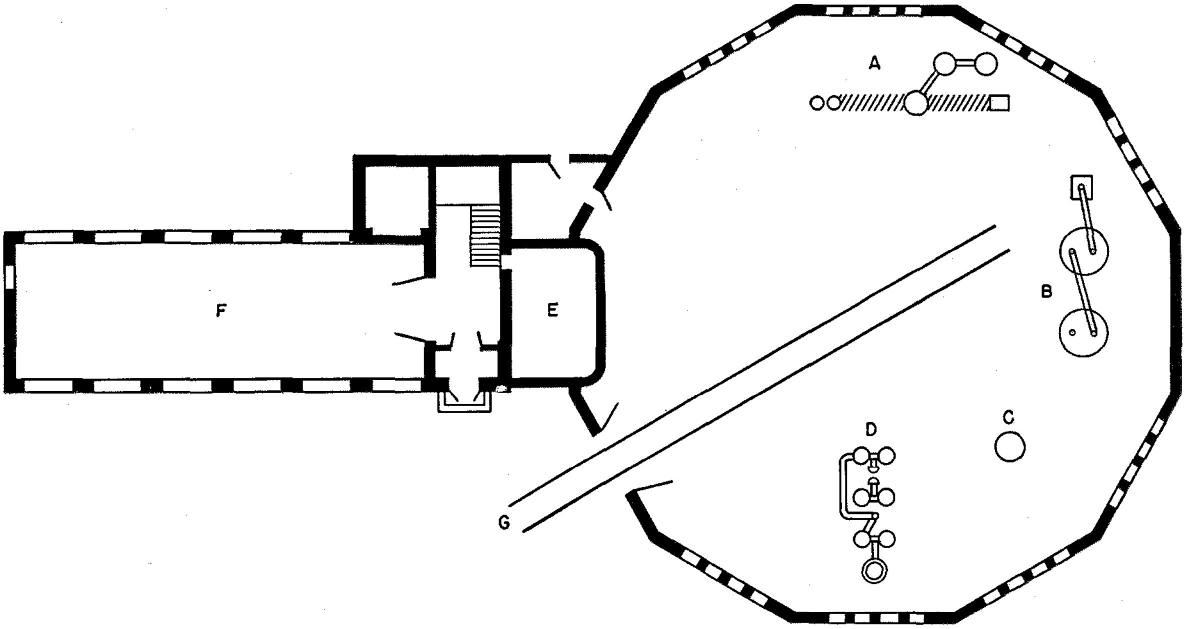


Figure 1—Plan of the laboratory. *A*—the 1 000 000-volt direct-current test set. *B*—600 000-volt alternating-current reactor set. *C*—1.5-meter-(4.9-foot-) diameter spark gap. *D*—2 100 000-volt impulse generator. *E*—the control room (on the second floor). *F*—physical laboratory, instrument shop, et cetera. *G*—railroad track.

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* Reprinted from *Electrical Engineering*, volume 74, pages 188–191; March, 1955.

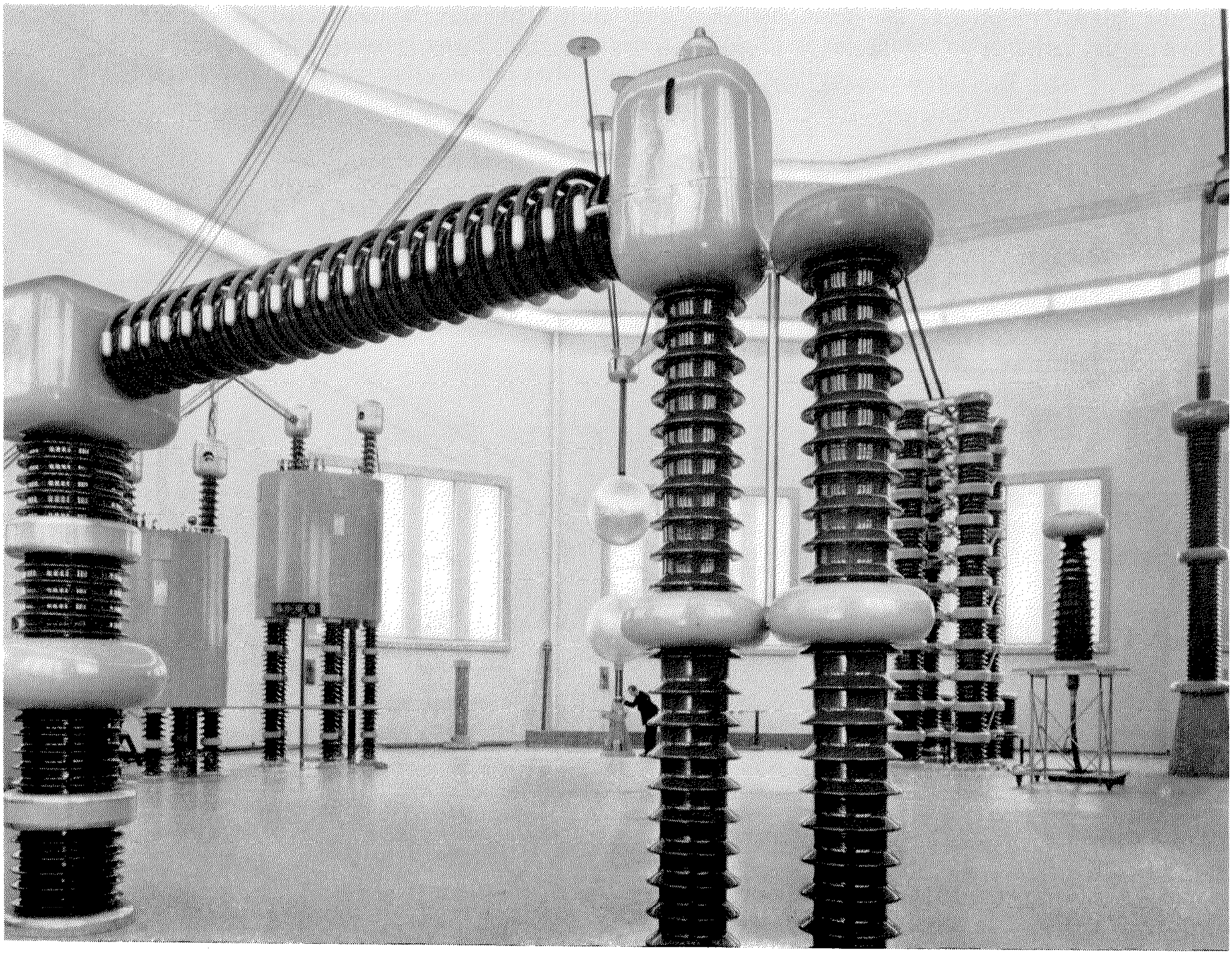
High-voltage alternating current for testing purposes is usually obtained by using cascade-connected transformers. With capacitive loads, as is usual in the cable industry, there is always a possibility of resonance between the cable and transformer reactances. If this should happen at one of the supply-frequency harmonics, the waveform of the terminal voltage will be distorted. Another disadvantage with the cascade principle is that the transformers do not divide the potential equally among themselves because of uneven distribution of the circuit reactance.

To avoid these difficulties, it was decided to substitute for the conventional transformers a series circuit (Figure 4) having variable inductances tuned to resonate with the cable capacitance at 50 cycles per second. The series principle has the additional advantage that the short-circuit current at breakdown is reduced to a fraction of the load current. This is very important, as it avoids burning up the test specimens after breakdown. The reactors are continuously variable and are operated from the control desk.

Figure 5 and the frontispiece show the surge generator, which is a conventional two-column Marx circuit for charging capacitors in parallel and discharging them in series. The circuit is shown in Figure 6. The highest charging potential is 175 kilovolts per stage, giving the 12-stage generator a maximum discharge voltage of 2100 kilovolts. The energy per charge is 30 kilowatt–seconds at full voltage. The generator has a low self inductance and is designed to give a standard surge with a 1-microsecond front and 50-microsecond tail on short cable lengths. There is sufficient headroom to increase the generator output to 2500 kilovolts by adding two more stages.

For recording the surge, there is installed a

Figure 2—The high-voltage laboratory.



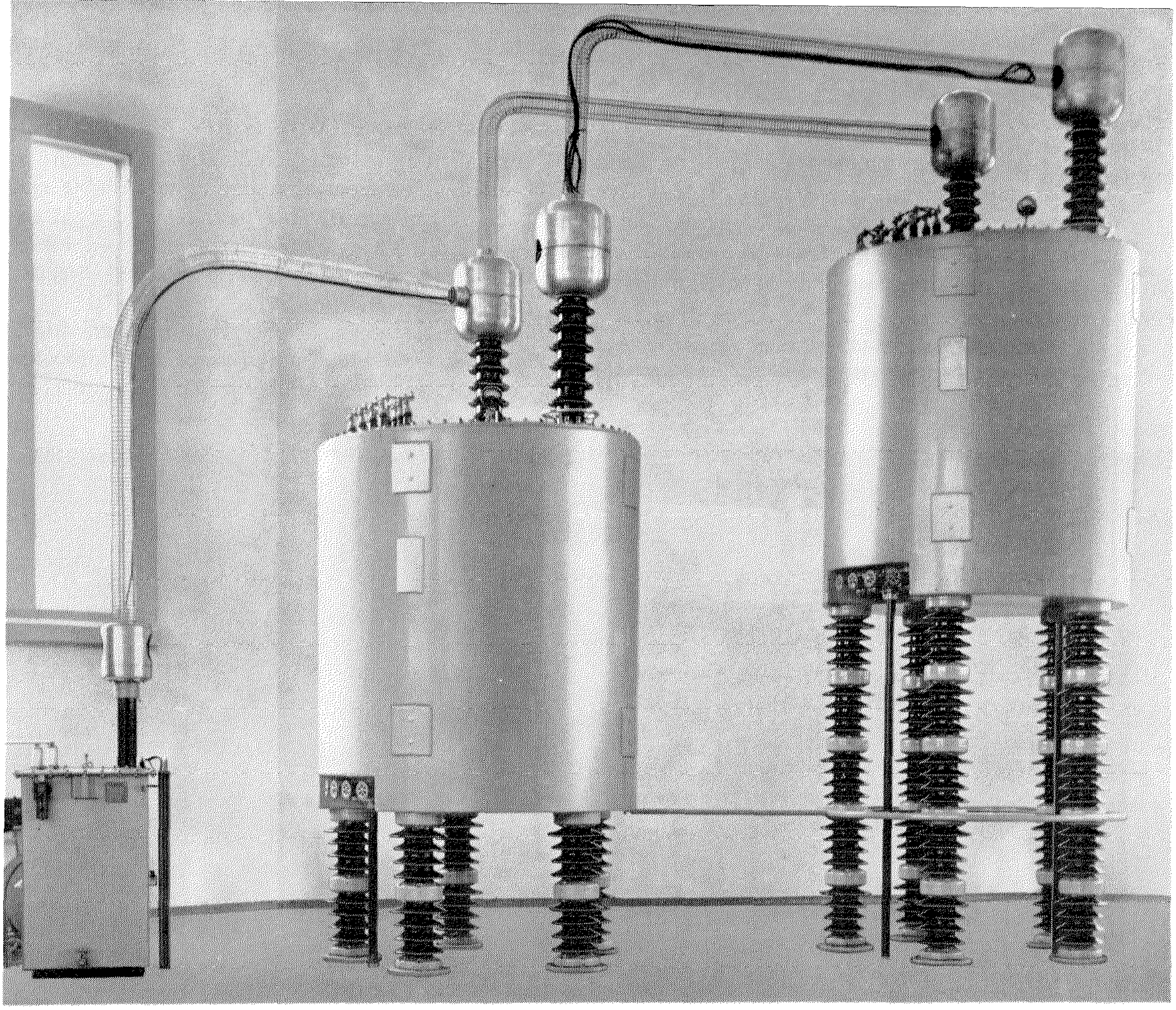
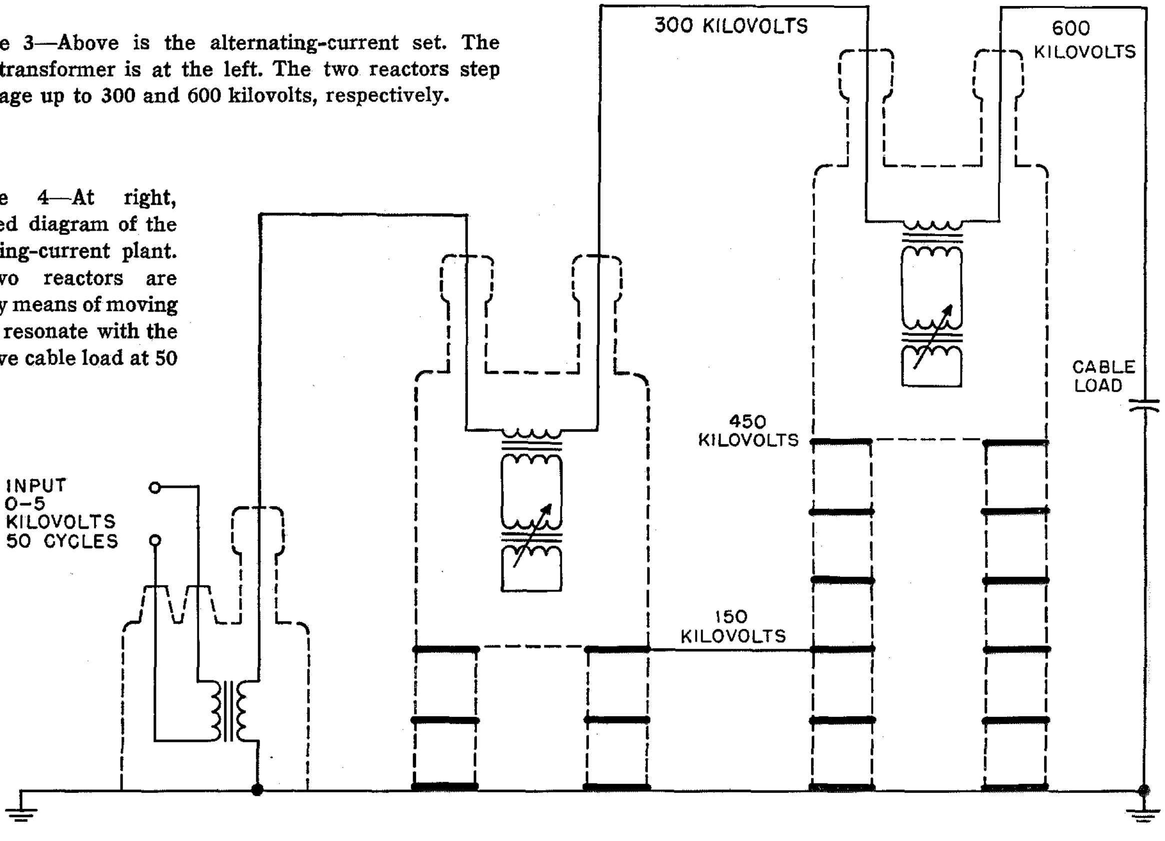


Figure 3—Above is the alternating-current set. The supply transformer is at the left. The two reactors step the voltage up to 300 and 600 kilovolts, respectively.

Figure 4—At right, simplified diagram of the alternating-current plant. The two reactors are tuned by means of moving coils to resonate with the capacitive cable load at 50 cycles.



two-beam continuously pumped oscillograph with an accelerating potential of 45 kilovolts, and having a writing speed of more than 6000 miles (10 000 kilometers) per second, or 10 meters per microsecond. The impulses are transmitted to the oscillograph through a coaxial cable from the foot of a variable capacitive voltage divider (seen to the right in Figure 5). The sweep can be observed on a fluorescent screen or it can be recorded directly on film.

Figure 7 shows the direct-current generator, which has a no-load potential of 1000 kilovolts and an output of 10 milliamperes. It is connected

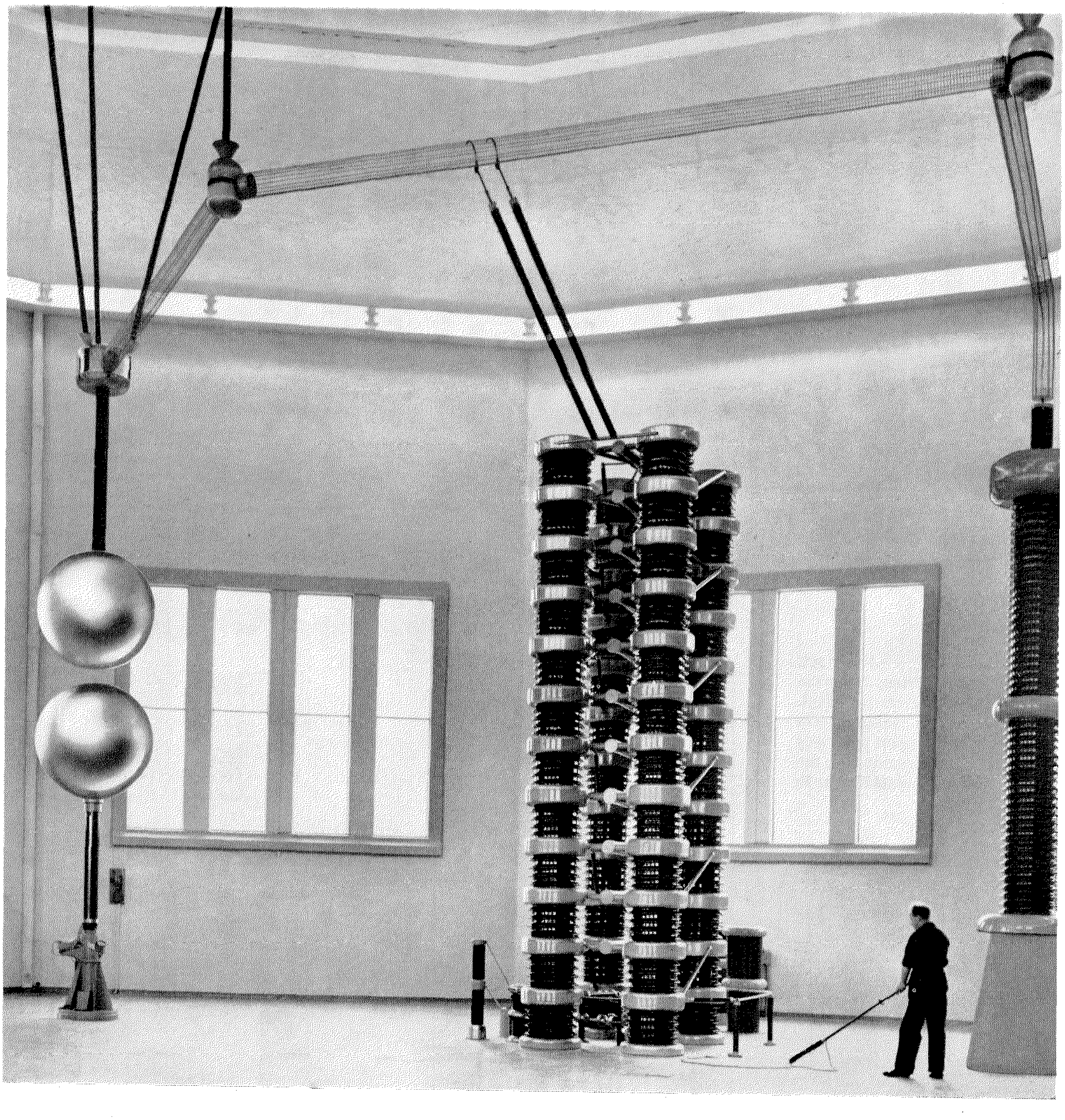
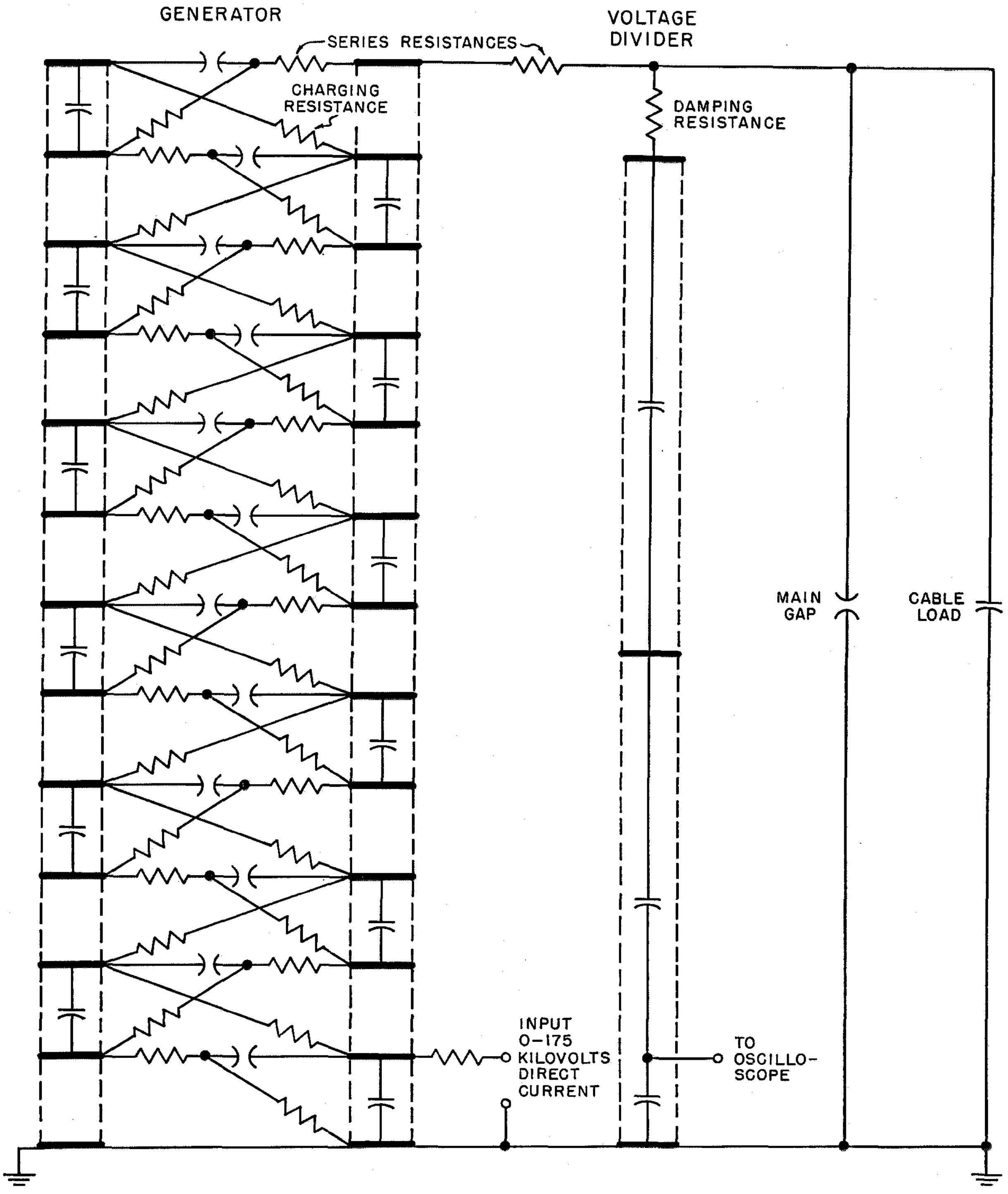


Figure 5—The 2100-kilovolt surge generator. From left to right are the spark gap, generator columns, and capacitive voltage divider.



as a single Cockroft–Walton voltage-doubler circuit (Figure 8). The horizontal member contains the rectifier consisting of 64 000 selenium cells. By means of external connecting links, the rectifier column can be broken into sections for parallel connection to give higher current

Figure 6—Schematic diagram of surge generator. The capacitors in the generator are 0.16 microfarad each. The capacitors are charged in parallel. When the gaps in the column break down, the capacitors are discharged in series across the load. The maximum surge voltage is set by the carefully calibrated main gap.

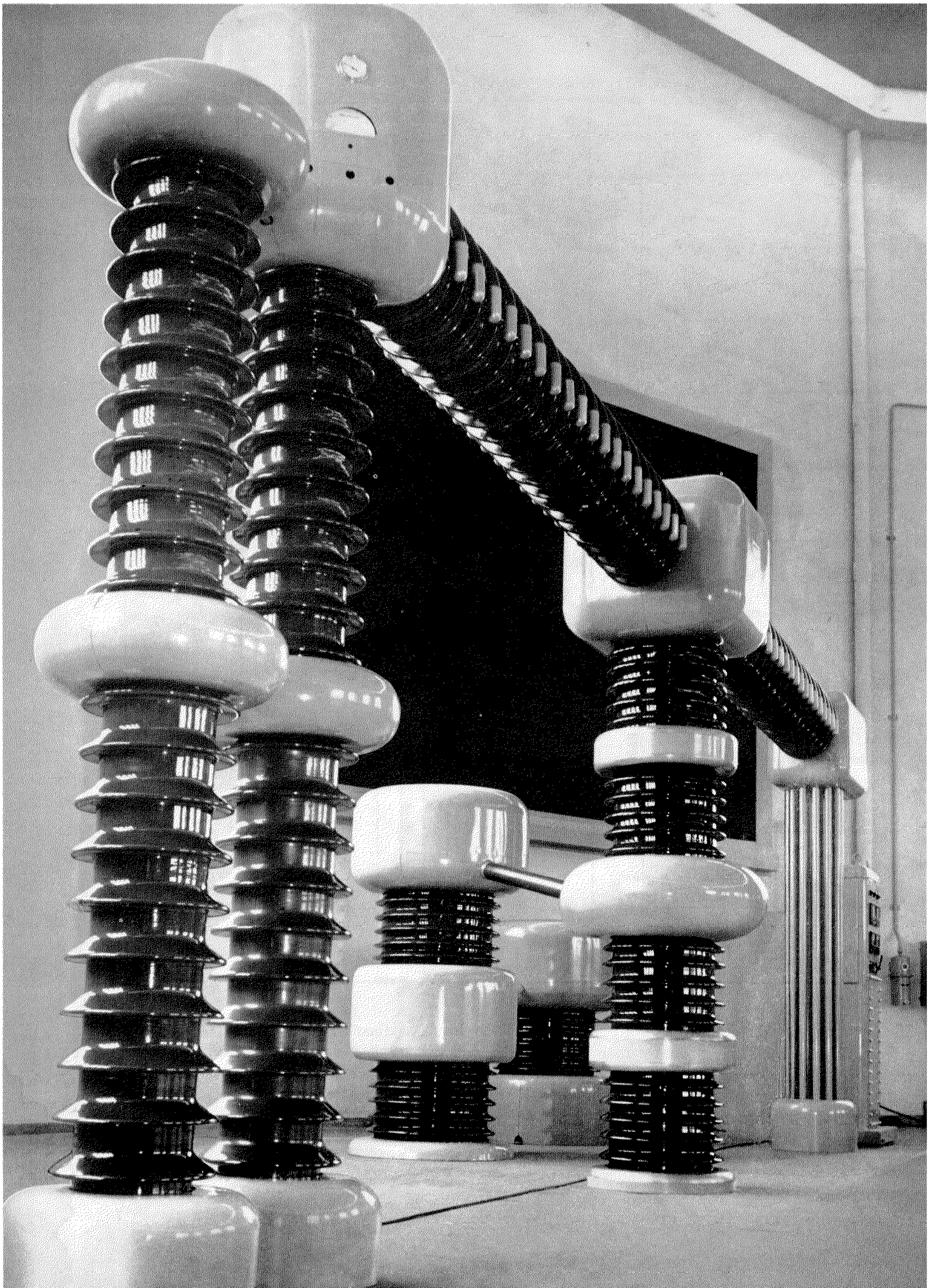


Figure 7—On the facing page is a view of the 1000-kilovolt direct-current generator.

capacities at lower voltages. The same arrangement also permits change of polarity. The generator is oil cooled, a circulating pump being mounted on the right-hand cooling column.

used for measurement of alternating voltages. In this case the voltage is read in kilovolts on a direct-current instrument.

A system of bus bars is suspended from the ceiling for distribution of the potentials to the various parts of the test area. An extensive sectionalized grounding system is accessible at a number of points through wells in the floor. A railroad track is laid across the test area, making it possible to test heavy pieces.

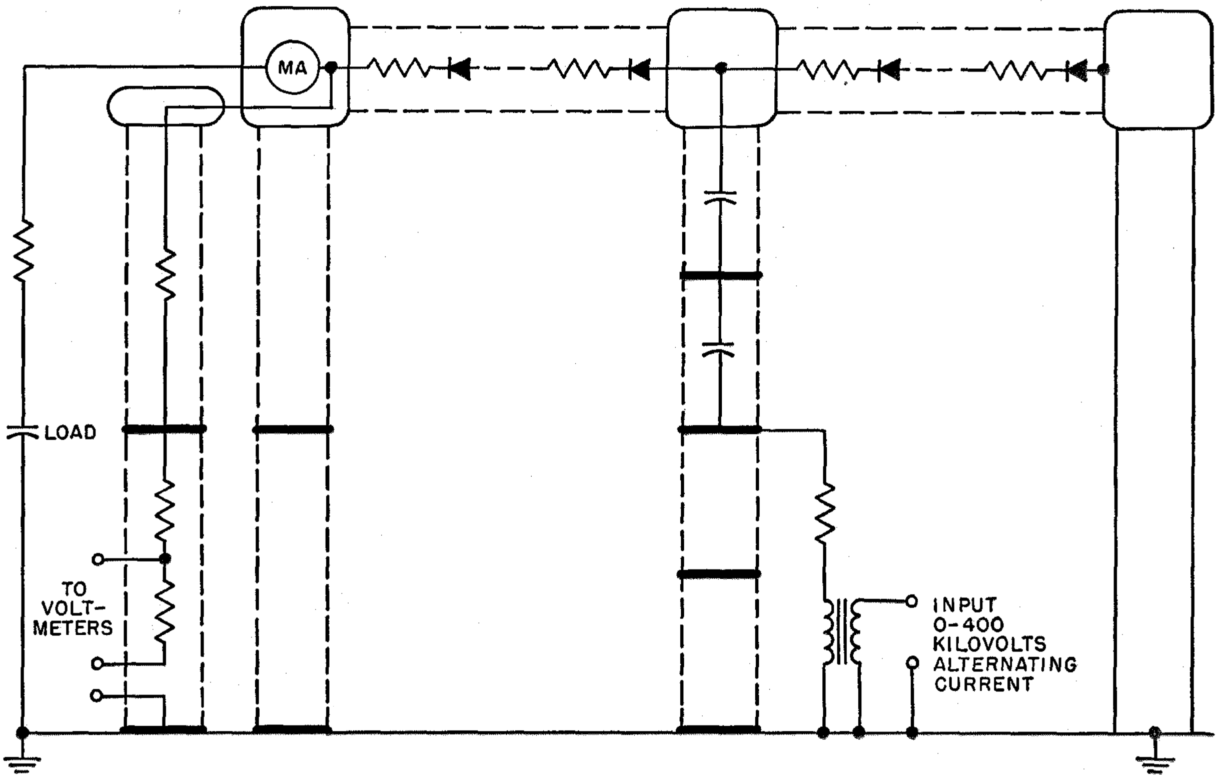
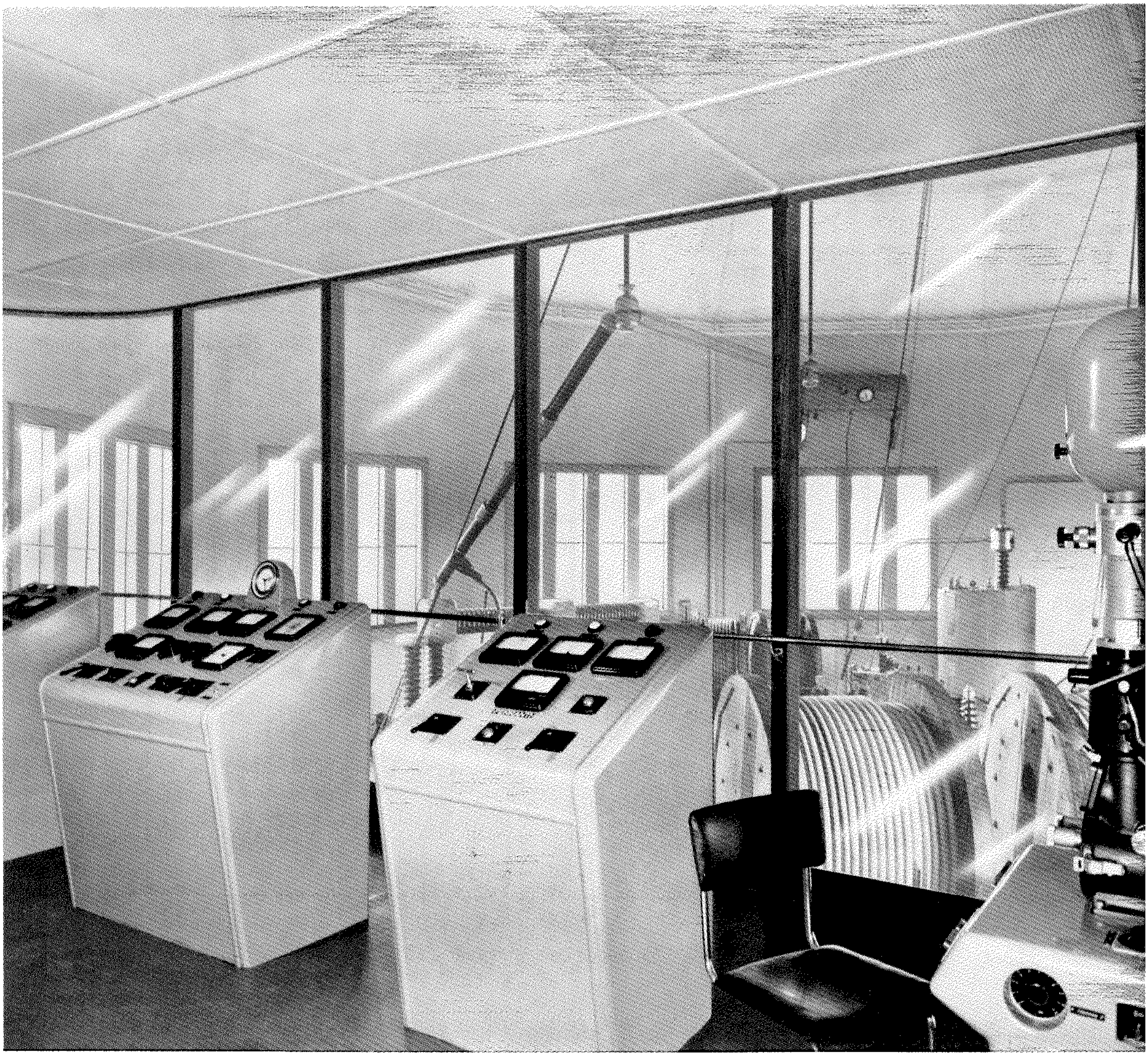


Figure 8—Circuit of the direct-current generator.

The control room is shown in Figure 9 with the oscillograph at the right. For checking high voltages and for calibration purposes, a spark gap with 1.5-meter- (4.9-foot-) diameter spheres is used. The gap distance can be adjusted remotely from the control desk, which also contains an instrument indicating the gap opening. As is well known, the spark-over voltage can be determined with good accuracy when the sphere diameter and the gap opening are known and adjustments are made for atmospheric conditions. The spheres are also

To prevent disturbance to nearby radio receivers, especially when the surge generator is in operation, the laboratory walls and roof are electrostatically screened.

With potentials of the magnitude involved, it is often of interest to study corona discharges. These discharges emit very feeble light and



must be observed in absolute darkness. For this purpose the main building is equipped with remote-controlled blackout curtains.

The high-tension testing equipment was delivered by Ferranti Limited; Manchester, England. The selenium elements were supplied by Standard Telephones and Cables, Limited of London and the oscillograph by Trüb, Taüber and Company A/G of Zurich, Switzerland.

Figure 9—The control room looks into the laboratory from the second-floor level. The high-speed oscillograph is at the extreme right.

Standard Telefon og Kabelfabrik A/S is the only Norwegian factory making high-tension cables; the new laboratory will make it possible for the company to advance with the technical development of this field.

Subminiature-Tube Reliabitity Testing

By STERLING BICKEL

Farnsworth Electronics Company, a division of International Telephone and Telegraph Corporation; Fort Wayne, Indiana

FAILURE of a vital element in a guided missile during flight can so reduce the performance and the information received from the missle as to make the experiment almost worthless. In many such instances, the unreliability can be blamed on vacuum tubes. Poor workmanship during construction and variations in characteristics with time and environment make tubes a major source of circuit failure.

Especially stringent requirements are imposed by tactical environmental factors, such as; vibration, shock, acceleration, temperature extremes, et cetera, and every electronic tube installed in equipment designed or developed for missile applications must be subjected to a rigorous test program. Some of the procedures followed in the reliability testing of subminiature tubes as they are received from the manufacturer are outlined in this article.

1. Mechanical Inspection

The first step in the test and inspection procedure is to count the tubes and to inspect them visually for damage to envelopes, leads, et cetera. They are next checked for envelope diameter. For a small shipment, this is a 100-percent check, while with larger shipments only a spot check is made. If any tubes fail the spot check, the entire shipment is checked for envelope diameter. This is an important factor in insuring a proper fit in the tube mount of wired-in subminiature tubes.

All tubes passing the mechanical inspection are marked with a serial number. This procedure permits correlation of subsequent failures with data obtained during incoming inspection, a practical method of establishing minimum qualities for acceptable tubes.

2. Electrical Tests

2.1 Transconductance and Plate Current

The transconductance of all tubes of the pentode and triode types is checked. At the same time, the plate current and screen current are checked. The operating potentials and inspection limits are those specified by the tube manufacturer. After these tests and while the tube is still in the bridge, the tube is checked for the change in plate current resulting from reducing the heater voltage by 10 percent. If the change in plate current is less than 10 percent of the value at rated voltage, the tube is acceptable; if the change in plate current is greater than 10 percent, the tube is a reject. All tubes rejected by this test are deleted from further tests and are marked and stored for later disposition.

2.2 Special Tests

Several special tests that are performed include gas checks according to military (*JAN*) specifications, test of the ratio of mutual conductance to plate current under the specific operating conditions at which the tube will be used, and keyed-rectifier and leakage tests. In addition, transconductance curves of screen-grid-to-plate and suppressor-grid-to-plate are run on request of the section that will use the tube.

Diode rectifiers are checked for emission and for voltage breakdown at the rated peak-inverse potential and at the rated heater-to-cathode potential. Thyratrons and voltage-regulator tubes are checked for electrical and vibration characteristics in the actual circuits in which they are to be employed. These and other special tests are performed only as required by the users.

3. Vibration Tests

In missile use, many tube failures result from vibration, so an effort is made to screen out noisy or microphonic tubes. All triodes and pentodes passing the electrical tests are subjected to a vibration test using an electrodynamic vibration machine. The tubes are vibrated at 40 cycles per second with an acceleration of 15 times gravity (15 *g*). The microphonic-noise output due to vibration is read on an oscilloscope. The tubes are then classified according to their peak-to-peak noise-voltage output during vibration, as shown in Table 1. Any intermittent shorts will show on the oscilloscope as large noise spikes and are cause for rejection of the tube under test. Tubes falling in classes 1, 2, or 3 are color coded as indicated in Table 1 and stored for use in missiles.

4. Microscopic Inspection

4.1 PURPOSE

The tests listed above have been performed on missile tubes for the past several years. However, the number of tube failures that have occurred during the tests of the missile electronic equipment has still been excessively high. In an effort to avoid some of these failures and to reduce the chance of missile failure in flight, a procedure has been set up for visually inspecting the tubes under a binocular microscope. By this process, it is possible to sort out many tubes that are likely to fail under the rigorous requirements of missile use.

4.2 BASIS FOR ACCEPTING OR REJECTING TUBES

In a vacuum tube, there can exist many internal faults that do not affect its basic electrical operation but that, if the tube is placed in a unit where vibration is an important factor, greatly impair the function of that unit. For example, such a tube may become microphonic or fail completely as the result of undesirable disposition of small bits of loose metal or insulating material within the envelope. Short-circuits or open-circuits may occur because of cracked insulators or defective welds. For these reasons, microscopic inspection is employed for segregating those tubes that are bad risks from the reliability standpoint. The inspection is made with a Bausch and Lomb type-*JK2* stereomicroscope, having two objectives and two sets of paired eyepieces with magnifications of 10, 15, 30, and 45. At present all subminiature tubes for missile use are being microscopically inspected for the following items.

4.2.1 Cathode Ribbons

A. A "mashed" weld, with or without discoloration, resulting in a serious reduction in thickness of ribbon material.

B. A twist at the weld that places the ribbon at right angles to the weld.

C. A sharp bend in the ribbon at the weld that results in a tear.

D. A sharp bend in any portion of the upper third of the ribbon adjacent to the weld.

E. Actual contact of the ribbon with the aluminum oxide coating of the filaments.

TABLE 1

SUBMINIATURE TUBE CLASSIFICATION CHART

Type	Load Resistance in Kilohms	Equivalent Cathode Resistance in Ohms	Plate Voltage	Screen Voltage	Limits of Noise Output in Millivolts Peak-to-Peak			
					1 Yellow	2 Red	3 Green	4 Reject
5636	10	150	100	100	40	120	120–168*	168*
5639	2	100	150	100	160	480	480–560*	560*
5718	10	150	100	—	20	60	60–70*	70*
5719	10	1500	100	—	10	30	30–70*	70*
5784	10	270	120	120	40	120	120–280*	280*
5840	10	150	100	100	30	90	90–168*	168*
5899	10	120	100	100	30	90	90–168*	168*
5902	2	270	110	110	160	420	None	420*
5977	10	270	100	—	20	60	60–140	140*
5987	2	2000	120	—	40	120	120–280*	280*
6021	10	100	100	—	20	60	60–140*	140*
6111	10	220	100	—	20	60	60–140*	140*
6112	10	1500	100	—	None	70	None	70*

* To be revised downward shortly; to be 2.8 times the manufacturer's latest root-mean-square noise limit.

F. A "taut" condition resulting from insufficient slack or too short a length of ribbon.

G. Insufficient contact area at the weld.

4.2.2 Welds in General

A. Broken or not welded.

B. Overwelding that produces either weld flakes or weld drippings.

C. Improper location of metals before welding.

4.2.3 Loose Particles in Envelope

A. Conducting materials usually from getter material or from weld flakes or drippings.

B. Nonconducting materials, such as mica flakes, dust, and aluminum-oxide heater insulation.

4.2.4 Getters

A. Underflashed, producing a tube that possibly could contain gas or could become gassy later.

B. Overflashed, leading to the presence of large pieces of getter material in the envelope.

4.2.5 Heater Insulation

A. Broken inside upper or lower edges of the cathode.

B. Heater-wire leads not properly pressed or welded in ribbons.

4.2.6 Micas

A. Cracked through.

4.2.7 Glass

A. Broken, cracked, or chipped.

4.2.8 Plates

A. Open seams in line of welds.

B. Burned from welding.

C. Splashed with weld or getter material.

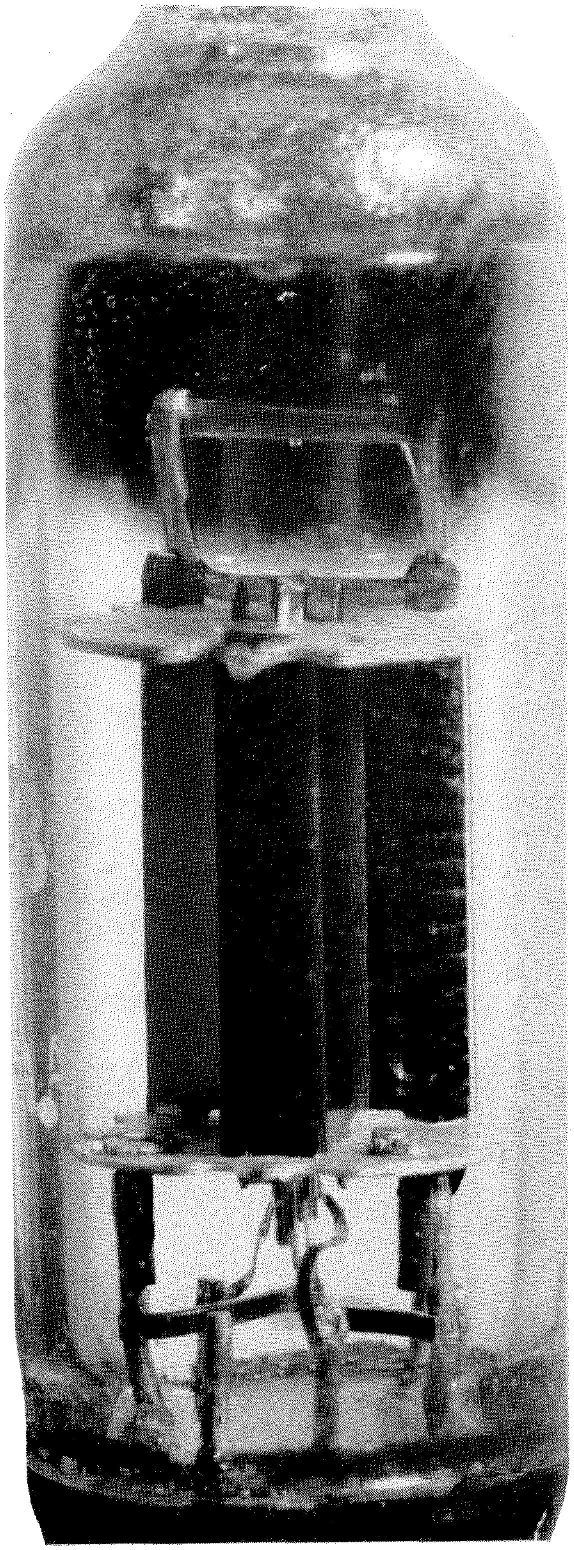


Figure 1—A good tube that may be used as a sample for the inspector.

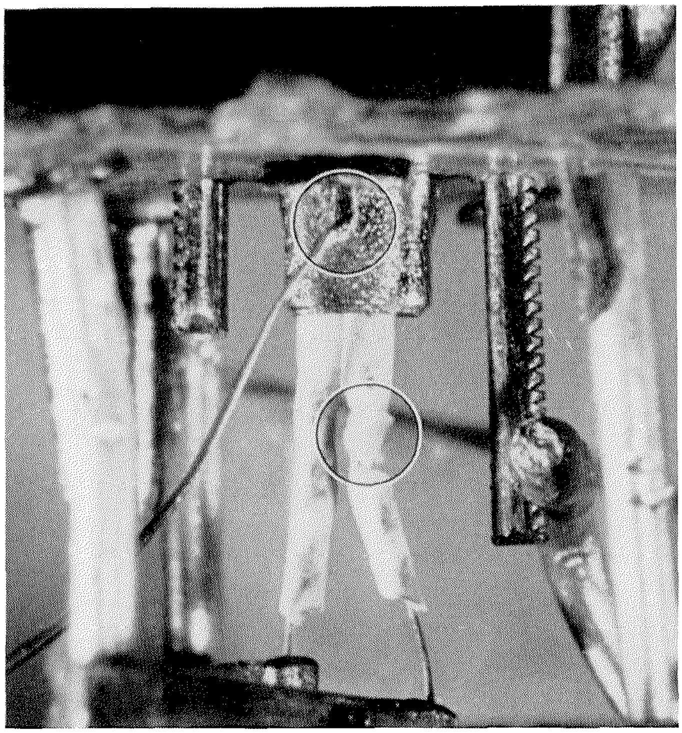
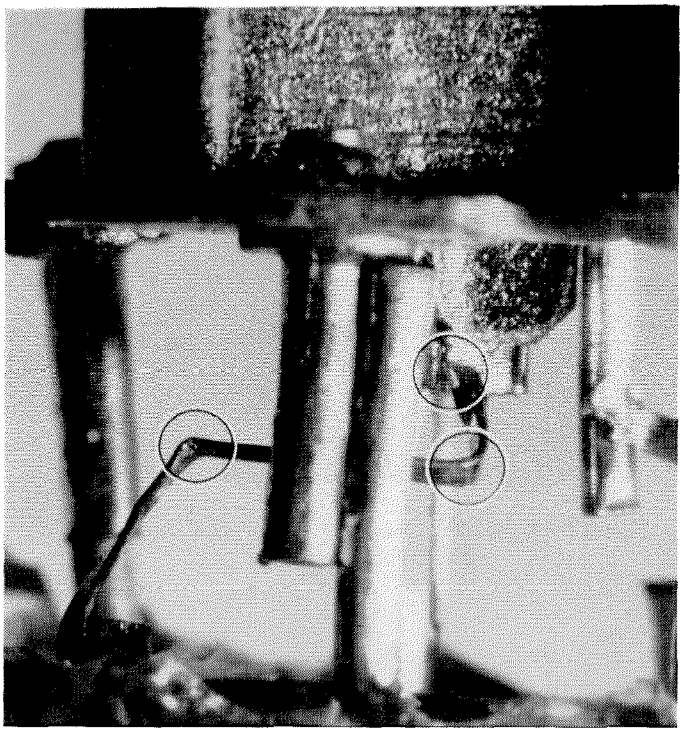


Figure 2—Broken heater insulation and a bent and torn ribbon lead to the cathode.

The eight major items listed above are in the order of their relative importance, with most attention being given during testing to the first six. Data have been taken and correlation work has shown that a definite relationship exists between tube failures and bad cathode ribbons, bad welds, particles, and overflashed getters.

Figure 3—Sharp bends in ribbons that often result in breakage during vibration.



No definite limits have yet been established for accepting or rejecting tubes by microscopic inspection. Some tubes can be readily determined to be "good" or "bad" just by observing the mentioned items under the microscope. Others will be borderline cases and can be classified effectively only by an experienced operator. An experienced operator developes an "engineering sense of sight" that enables him to separate the good tubes from the bad; it is believed that once he has achieved this attitude, tube failures in the field can be materially reduced.

At present, tubes with obvious defects are being rejected. Most of these have defective cathode ribbons, since this has been found to be one of the most serious faults encountered in the field. Other defects of a minor nature will be recorded against the serial number of the tube, and field failures will then be correlated with this inspection record.

5. *Examples*

Figure 1 demonstrates the appearance of a "good" tube that is suitable for the establishment of a standard.

Figure 2 reveals broken heater insulation and a bend and tear in the cathode ribbon at the cathode. Faulty heater insulation such as is shown in Figure 2 will possibly break off in service and become a noise generator. The bend and tear in the cathode ribbon at the cathode weld is cause for rejection, since in the past this has proved to be a major cause of tube failure.

Figure 3 illustrates some of the sharp bends that commonly occur in cathode ribbons. Many times the ribbons are actually kinked at these bends. Such bends cause the ribbon to break quickly during the cycling process.

An example of a bad cathode weld is shown in Figure 4. The weld has been mashed and overheated, and has a bad twist. It has been determined that such flaws at this junction will cause failure during the early part of what would be the normal life of the tube.

Figure 5 is an illustration of a too-tight cathode ribbon. In some tubes the tautness of the ribbon has resulted in a complete break some-

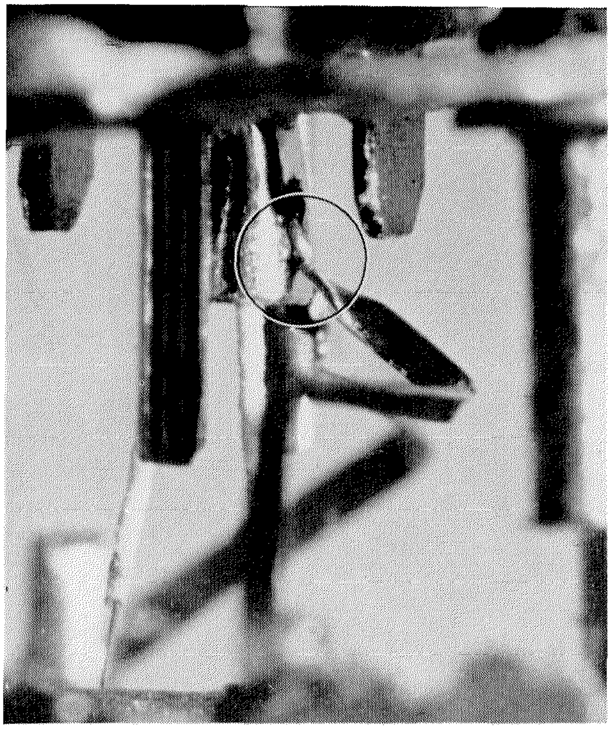
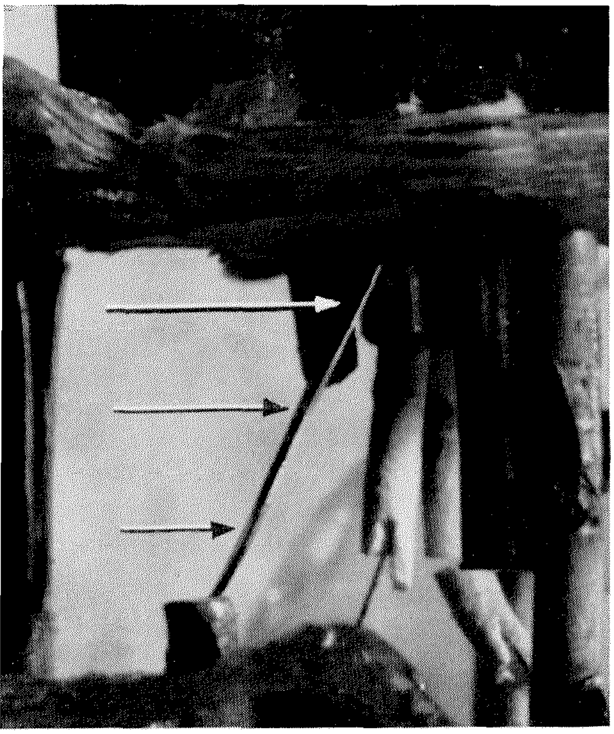


Figure 4—A bad weld.

where in the upper third of its length. Such breaks result from the expansion and contraction of the ribbon with repeated heating and cooling of the cathode. As temperature changes in the cathode take place, the ribbon must move with a motion similar to that of the human elbow.

Figure 5—Cathode ribbon pulled too tight.



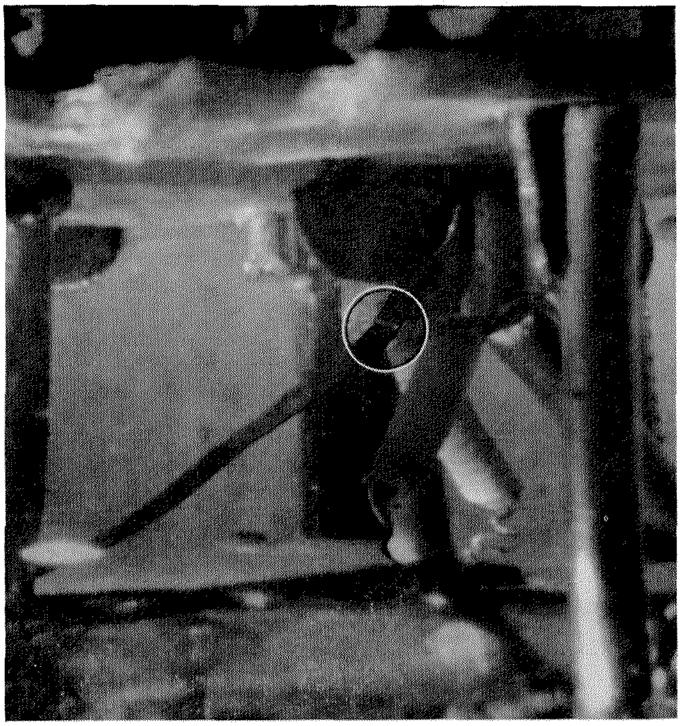
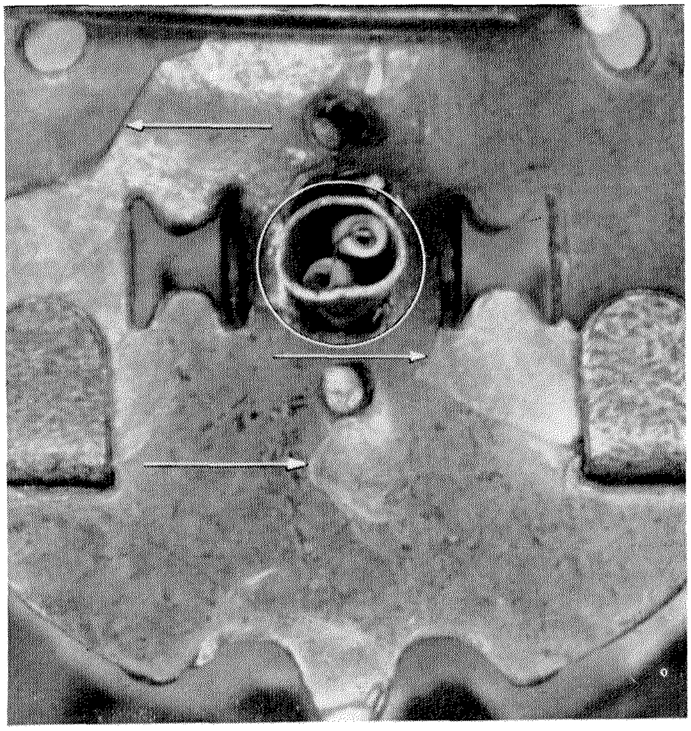


Figure 6—This overly tight cathode ribbon has broken after a few hours of testing.

A small distance below the cathode in Figure 6, a break in the cathode ribbon can be seen. This break occurred during test after a few hours of operation. The break occurred as a result of a taut ribbon condition and could have been detected by microscopic inspection.

Figure 7—Broken insulation at the top of the heater.



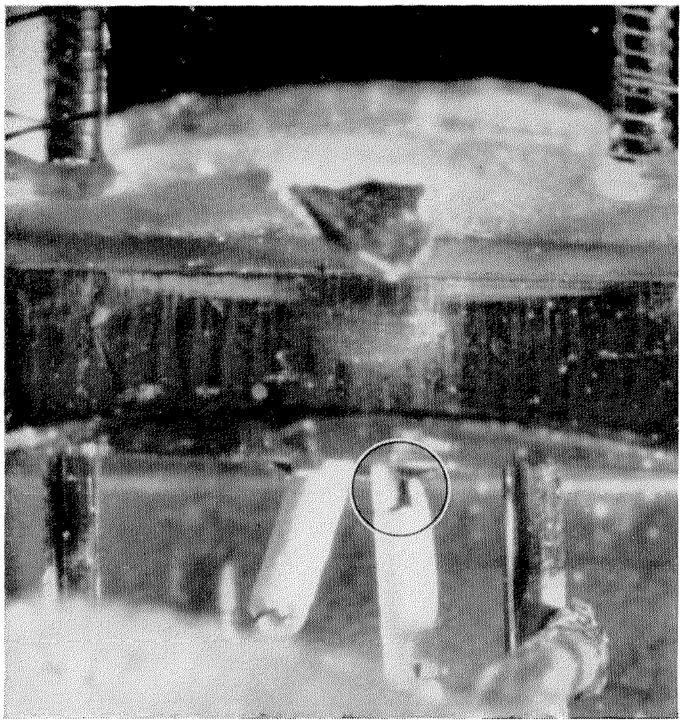


Figure 8—Broken insulation at the bottom of the heater.

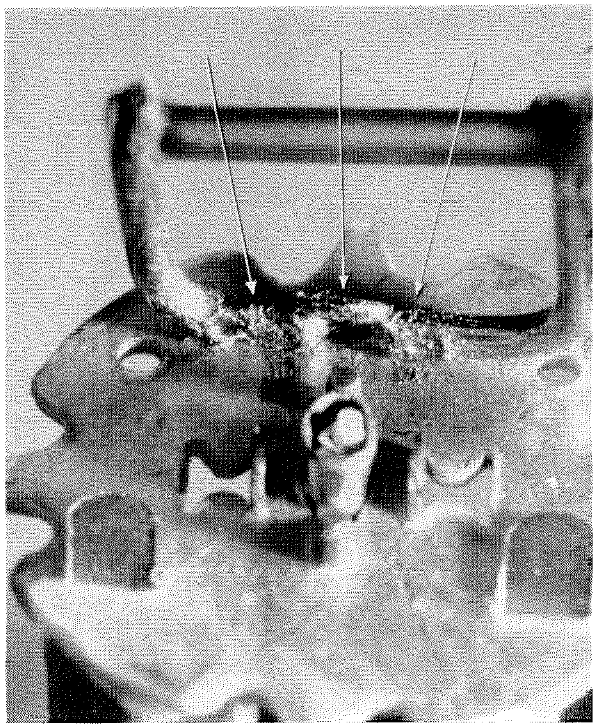


Figure 10—Bad weld that has burned through the getter support.

Figure 7 shows broken heater insulation at the top of the cathode. The heater insulation of this tube is broken down inside the upper part of the cathode resulting in a very bad heater-to-cathode short. This condition is due to poor workmanship in coating the heaters or in installing them in the cathode. Note also the large flakes of mica beginning to peel from the upper mica support.

Figure 9—This getter strip has been overheated and the burned-out piece will cause noise.

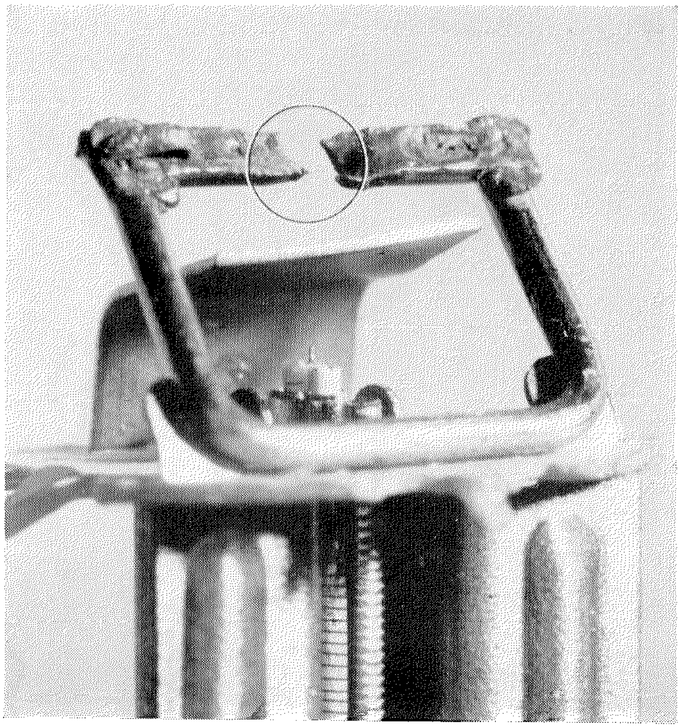


Figure 8 reveals a break in the heater insulation at the lower edge of the cathode. This break caused a short in the tube pictured and, if it had not been detected in time, could have caused the cathode ribbon to burn in two.

Shown in Figure 9 is an example of getter overflashing resulting from too much induction heating in the flashing process. With overflashing, there is always present a large piece of metallic getter material, either loose in the envelope or clinging to one of the elements. In the latter case, experience has shown that with service such particles break loose and cause a tube to become noisy.

Notice in Figure 10 the extremely bad weld between the getter and the plate; the getter support is burned apart. As a result of the poor weld, many spattered clinging particles are present in the envelope.

In Figure 11, a piece of loose getter material can be seen lying just below the lower mica and against the support rod of the third grid. During vibration, the motion of this piece of material

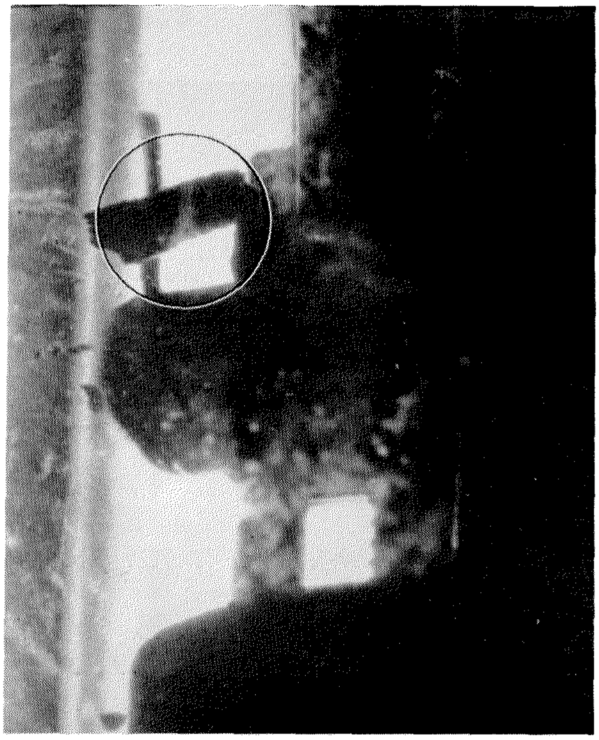


Figure 11—Piece of loose getter material.

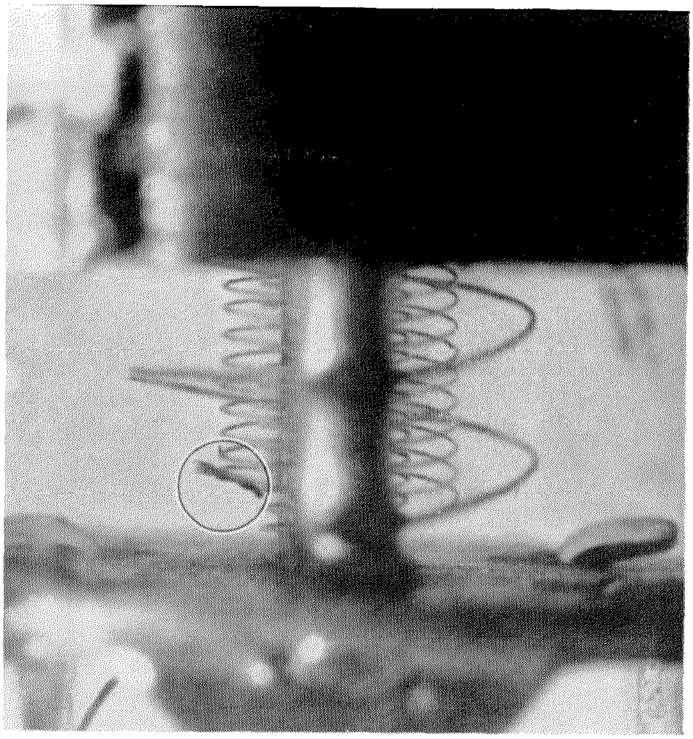


Figure 12—A piece of material short-circuiting the first and second grids.

inside the envelope of the tube will cause noise or intermittent shorts.

Close inspection of Figure 12 will reveal a piece of getter material or weld flake lying between the third turn of the inner-grid winding and the first turn of the second-grid winding, resulting in a direct short-circuit between these two grids. Such a condition is brought about by poor workmanship during production of the tube.

Lying against the envelope of the tube shown in Figure 13, there can be seen a number of particles of heater insulation, weld flakes, and getter material. Some of the particles are metallic and some are nonmetallic. Particles of this kind are a potential noise source.

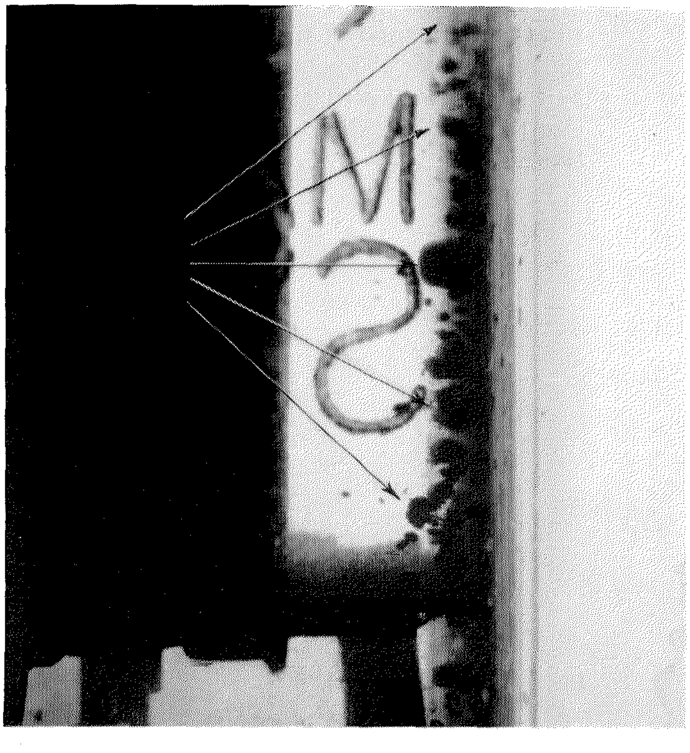
Close observation of Figure 14 will reveal the presence of a piece of weld splash clinging to the weld at the lower end of the grid-support rod. During vibration tests, this metallic particle would probably break loose and cause noise or a short. Note, also, in Figure 14, the taut condition of the cathode ribbon.

Figures 1 through 14 illustrate some of the worst conditions encountered in tubes. Visual microscopic inspection can keep a large percentage of tubes having these faults out of finished equipment.

6. Desired Achievements

The ultimate goal of visual microscopic inspection is to eliminate as many as possible of the tubes having the worst conditions described

Figure 13—Collection of conducting and insulating particles within a tube.



above. This is now being accomplished by means of incoming inspection. In the past, numerous tube failures in the field could have been detected with a microscope. Of course, microscopic inspection is not the final answer; the real solution to the problem can be reached only by obtaining the manufacturers' cooperation in cleaning up their production lines. This will take considerable time, however, and until it can be accomplished, visual microscopic inspection will be a useful tool.

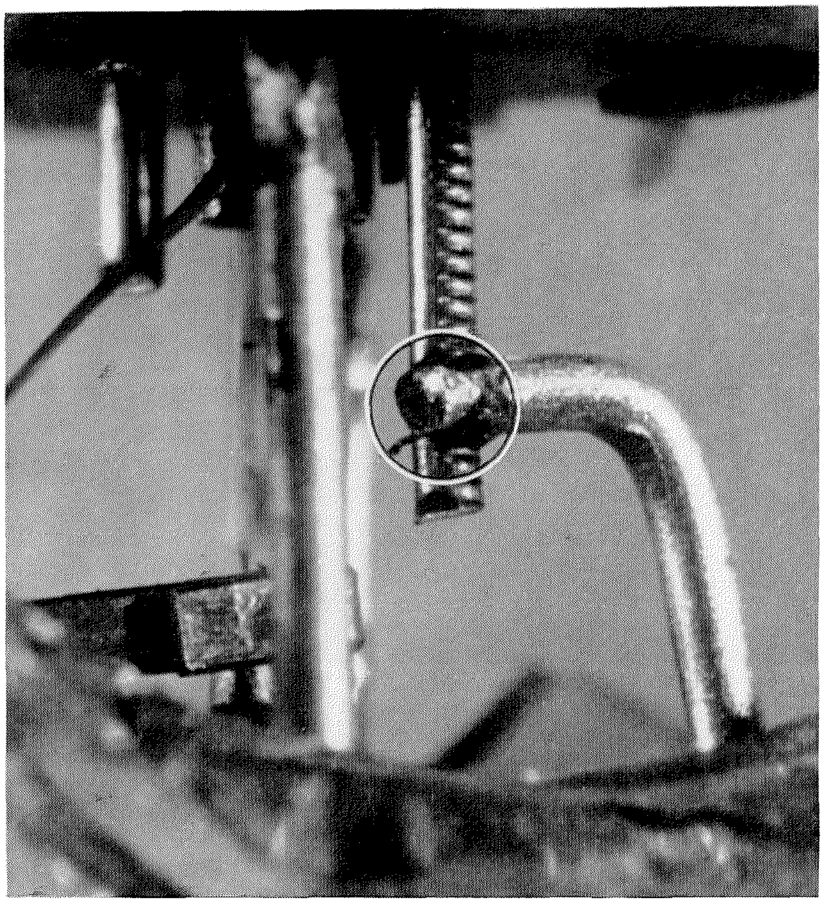


Figure 14—Piece of weld splash that may break away under vibration and cause noise. Also, the cathode ribbon is too tight.

The entire tube inspection program is necessary because even the rigorous quality control practiced by tube manufacturers is insufficient to satisfy the exacting requirements of missile use. It is hoped that at some future date it will be necessary for the user to make only spot checks in order to assure satisfactory quality, with the main burden of inspection being assumed by the tube manufacturer. Until that time, however, the microscopic 100-percent incoming inspection, as well as the described mechanical inspection, electrical tests, and vibration tests, will be continued.

One duty will always rest with the user, that is, the duty to keep adequate records of the failures that occur in actual use. Only by reporting such failures and returning the defective tubes to the proper agencies for analysis can the causes of failure be ascertained and corrected.

Electromagnetic Delay Lines*

By H. G. NORDLIN

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TWO BASIC KINDS of devices are used for providing known amounts of phase shift or time delay of electrical signals. With supersonic delay devices, time delays from several microseconds to several milliseconds can be achieved. Their efficiency is, however, greatest for large time delays because of the large fixed insertion loss of the electromechanical transducers. For time delays of a fraction of a microsecond to about 10 microseconds, electrical delay devices are generally perferred because they do not have fixed insertion losses and the attenuation for a given design is proportional to the amount of time delay. In addition, the electrical delay devices are comparatively insensitive to mechanical shock and changes in temperature, which often give trouble in the supersonic delay devices.

An ideal delay line would be one in which the shift in phase from input to output is proportional to frequency and in which the voltage attenuation is a constant for all frequencies of interest. Any input wave, provided it has no components outside the prescribed frequency range, would be transmitted to the output of the ideal device without distortion. These ideal conditions cannot be realized in practice, and deviations from them are therefore a measure of the fidelity of a delay line.

Electrical delay lines in present use are of two types; lumped parameter and distributed parameter. In the usual cases, both are 4-terminal networks composed of series inductance and shunt capacitance. Resistance and dielectric dissipation are inherent properties of the inductors and capacitors, respectively. The simplest type of lumped-parameter delay line is the contant-k low-pass filter. This device exhibits a time delay characteristic that begins to increase with frequency about a decade below the cutoff frequency, which is directly proportional to the number of T sections in a line of a given time delay and impedance. The variation of time delay may be reduced or even reversed in direction below the cutoff frequency by causing the adjacent inductance sections to be mutually coupled to an increasing degree.

If the number of sections in the low-pass constant-k delay line is increased without limit, we obtain a line in which the parameters of inductance and capacitance are distributed uniformly between the input and output terminals. This resulting distributed-parameter delay line does not exhibit the sharp cutoff found with the lumped-constant devices. The useful frequency limit is approached rather slowly with increasing frequency and is evidenced by an increase in both velocity and attenuation. In general, distributed-parameter delay lines can be designed to have a greater useful bandwidth than lumped-parameter lines of equivalent size and time delay.

Our recent work in this field has been in the development of distributed-parameter delay lines of the cable type that are manufactured in continuous long sections. Some delay cables that were developed under contract with Squier Signal Laboratories of Fort Monmouth, New Jersey, will be described and data will be given on their response to rectangular pulse and sinusoidal waveforms.

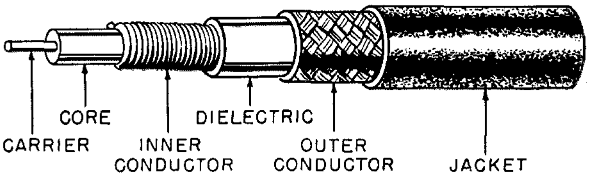

Figure 1—Distributed-parameter delay cable.

Figure 1 shows a cutaway sketch of a distributed-parameter delay cable of the type that is familiar to many electronic engineers. It consists of a flexible inner core over which are disposed the inner conductor of insulated wire wound in the form of a closely spaced single-layer solenoid, a dielectric layer that may be extruded or formed with a tape, an outer conductor that may be either a braid or serving of insulated or bare copper wire, and a vinyl-type jacket that acts to prevent loosening of the outer conductor and also as a protection against physical damage and moisture. An outstanding advantage of the distributed-parameter delay cable lies, of course, in the fact that a required amount of time delay can be obtained merely by cutting off a piece of cable of the necessary length.

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* Reprinted from *Proceedings 1954 Electronic Components Symposium,* pages 153–157. Presented at the Electronic Components Symposium in Washington, District of Columbia, on May 5, 1954.

The original cable of this type, *RG-65/U*, was developed from a design originated at the Massachusetts Institute of Technology. It was intended as a high impedance cable, but since it has a time delay of 0.04 microsecond per foot (0.13 microsecond per meter), about 27 times that of a polyethylene-insulated coaxial cable, it has been used frequently as a broad-band distributed-parameter delay cable.

Three recently developed types, *A*, *B*, and *D* will be referred to in this paper. Types *A* and *B* were designed for highly stable operation over a wide frequency range at impedance levels of 1000 ohms and 250 ohms, respectively. Special manufacturing techniques were devised to insure a high degree of precision and uniformity in their production. The *D* cable is currently in an advanced stage of development. It represents the results of an effort to obtain high efficiency with regard both to voltage attenuation and concentration of time delay, as well as a reasonably high fidelity up to a frequency of 10 megacycles per second.

The *A* cable has a nominal time delay of 0.2 microsecond per foot (0.7 microsecond per meter) and a nominal characteristic impedance of 1000 ohms. Its overall diameter is 0.405 inch (10.3 millimeters). The inner conductor is size-32 (American Wire Gage) type-*F*-Formex-coated magnet wire, and the outer conductor is a conventional braid of size-34 *F*-Formex-coated magnet wire. The response of the *A* cable to rectangular pulse and sinusoidal waveforms is illustrated in Figure 2. The time delay is constant from 1 to 10 megacycles per second and decreases by about 5 percent from 10 to 30 megacycles. The attenuation of the cable is 3 decibels per microsecond of delay at about 10 megacycles.

The oscillograms show how a rectangular pulse is distorted after passing through 6 feet (1.8 meters) of cable terminated in its characteristic resistance. The picture on the left shows the shapes of the input and output pulses as well as their relative positions on a time scale. In this case, the delay is about 1.3 microseconds. The picture on the right shows the leading edge of the output pulse with the time scale expanded to 0.1 microsecond per division. The rise time of the pulse between the 10- and 90-percent points is 0.06 microsecond.

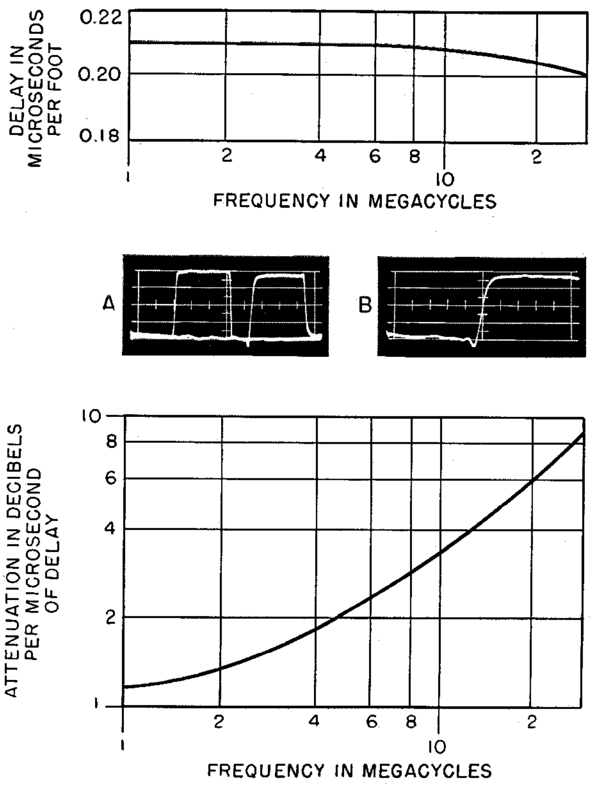

Figure 2—Type-*A* delay cable. The delay–frequency characteristic is shown at top and the attenuation–frequency relation at the bottom. Oscillograms are for a 1-microsecond pulse through 6 feet (1.8 meters) of cable with timing divisions of 0.3 microsecond at *A* and 0.1 microsecond at *B*. At *A*, the input pulse is at the left and its output form is at the right.

Similar information is shown for the *B* cable in Figure 3. This cable has a nominal time delay of 0.1 microsecond per foot (0.3 microsecond per meter) and a nominal characteristic impedance of 250 ohms. Its over-all diameter is 0.405 inch (10.3 millimeters). The inner conductor is size-23 (American Wire Gage) *F*-Formex-coated magnet wire and the outer conductor is size-34 *F*-Formex wire. The variation of time delay in the frequency range shown is about 5 percent and the attenuation curve is nearly the same as for the *A* cable. The rise time of the pulse, delayed through 10 feet (3 meters) of cable, is 0.06 microsecond.

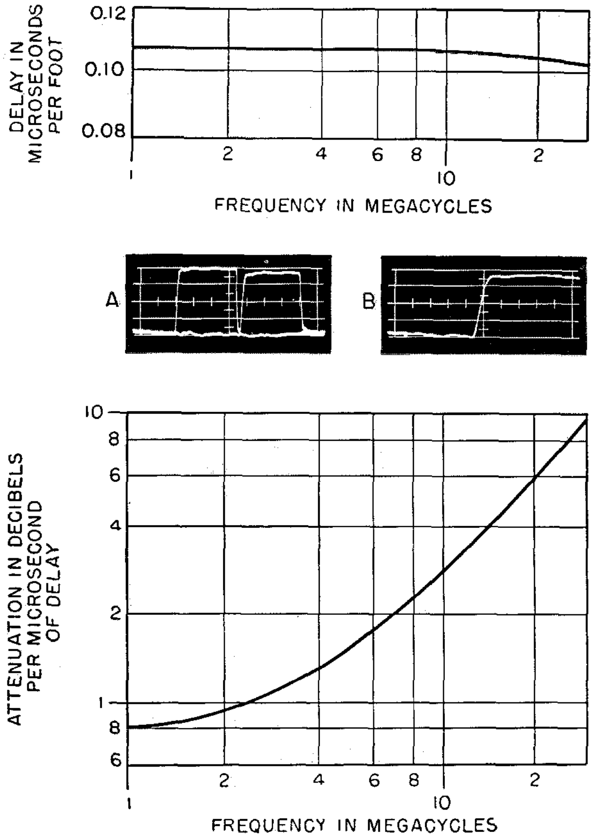

Figure 3—Type-*B* delay cable. The oscillograms are for a 1-microsecond pulse transmitted over 10 feet (3 meters) of cable with timing divisions of 0.3 microsecond at *A* and 0.1 at *B*.

In the course of the development work on the *A* and *B* cables, investigations were made of the magnetic properties of plastic compounds containing powdered ferrite materials with a view toward employing them in the core structure of the delay cables for the purpose of obtaining a lower attenuation per unit of time delay. We concluded that the losses introduced by the magnetic compound would overshadow the reduction in conductor losses at frequencies above about 10 megacycles but that a considerable improvement in efficiency could result in the 10-megacycle range. In addition to the possibility of improved efficiency, the use of magnetic materials makes it possible to increase the amount of time delay in a given volume of cable. Because the *A* and *B* cables were designed for the highest possible operating frequency, no magnetic materials were used in their construction.

We are presently in the final stage of the development of a *D* cable utilizing the desirable properties of ferrites. The objective is a cable of high quality having a nominal time delay of 1 microsecond per foot (3 microseconds per meter) length, low attenuation, and minimum amplitude and phase distortion in the frequency range up to 10 megacycles. Referring momentarily back to the data on the *A* and *B* cables, it will be recalled that the time delay on those cables is constant up to at least 10 megacycles, decreasing slowly at higher frequencies. This drop-off of time delay with increasing frequency has been shown in the literature to be due to the nature of the magnetic coupling between each turn of the inner conductor and neighboring turns. At low frequencies, the fields coupled in this manner are substantially in phase and the mutual coupling tends to enforce the general magnetic field. As the frequency is increased, coupling occurs to an increasing extent between turns in which the currents are not in phase and cancellations occur in the general magnetic field with a resulting decrease in inductance. It can readily be seen that for a greater concentration of time delay or, in other words, a larger phase constant, the out-of-phase coupling will occur to an important extent at a lower frequency.

In the *D* cable, the central core is composed of a thermoplastic compound containing a commercial ferrite in powder form. The core is extruded over a Fiberglas cord at a rate somewhat slower than is normally used for polyethylene. It has good flexibility and does not crack or break as a result of normal handling, although breakage will occur if the core is repeatedly bent in a radius of less than about 0.5 inch (12.7 millimeters). The inner conductor is size-40 *F*-Formex-coated magnet wire. Over this, a dielectric tape is wrapped with a 50-percent overlap. The outer conductor is a braid of size-36 *F*-Formex-coated magnet wire. The over-all diameter of the cable including the jacket is approximately 0.25 inch (6.4 millimeters).

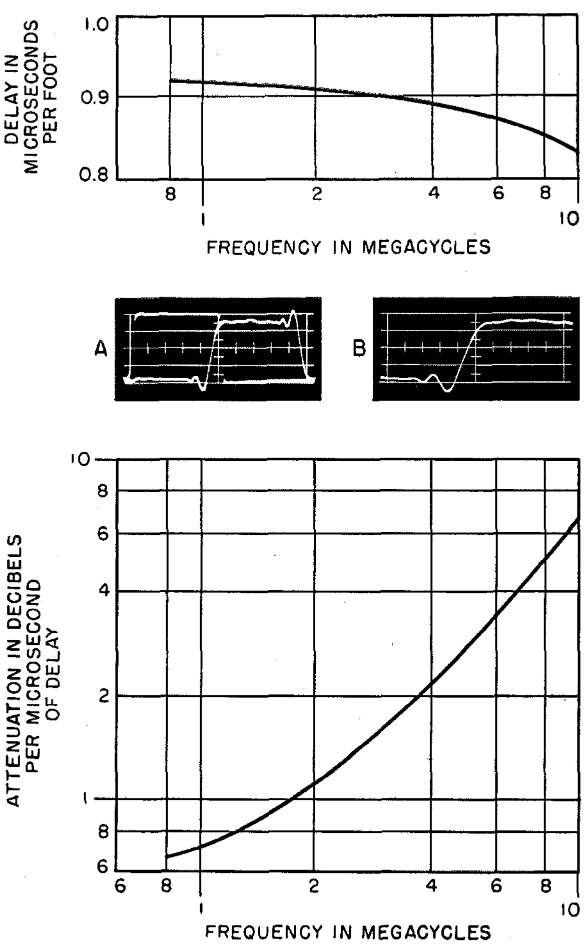

Figure 4—Type-*D* delay cable without compensation. The oscillograms are for a 1-microsecond pulse transmitted over 1 foot (0.3 meter) of cable with timing divisions of 0.2 microsecond at *A* and 0.1 at *B*.

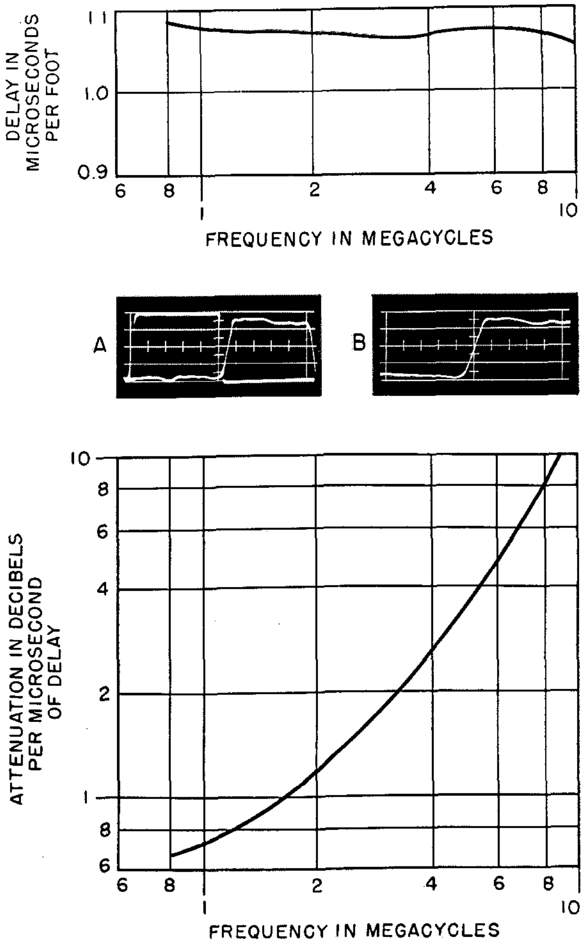

Figure 5—Type-*D* delay cable with metallized-film compensation. The oscillograms are for a 1-microsecond pulse through 1 foot (0.3 meter) of cable with timing divisions of 0.2 microsecond at *A* and 0.1 at *B*.

The response of the basic *D* cable to sinusoidal and rectangular pulse waveforms is shown in Figure 4. Whereas the *A* and *B* cables have a constant time delay between 1 and 10 megacycles, the time delay of this cable decreases about 10 percent in that frequency range. Although the attenuation at one megacycle per second is lower than for either the *A* or *B* cables, the increase at higher frequencies is greater, with the 3-decibel-per-microsecond point occurring at about 6 megacycles. The effect of phase distortion on the pulse response is seen to be the occurrence of oscillations or ringing preceding the 2 step changes in voltage on the pulse. In addition, the rise time of the pulse has increased to 0.12 microsecond after transmission through a 1-foot (0.3-meter) length of the cable. The nominal characteristic impedance of this basic *D* cable is 1750 ohms.

A method of compensating for the decrease in time delay of distributed-parameter devices by the use of floating metallic "patches" that are insulated from the helical inner conductor, from each other, and from the outer conductor, was described by Kallmann.[1] The decrease in time delay at high frequencies resulting from a decrease in inductance is thereby compensated for by a rise in the effective capacitance. Kallmann demonstrated the method by using rectangular pieces of metallic foil that were inserted one by one under the inner conductor during the construction of a delay-line section. It was shown that the time delay can be made nearly constant up to a frequency at which the patches are each about a half wavelength long. The degree of equalization is a function of the width of the patches and of the dielectric constant and thickness of the insulation between the patches and the inner conductor. The capacitance patches have no effect at low frequencies, where negligible phase difference exists along the length of each patch. As the frequency is increased and the phase shift increases, current flows through the patches and the effective capacitance of the cable is increased. This technique would suffice for the manufacture of small units but it would be difficult to devise a means of automatically installing such patches on a cable-type delay line in factory production.

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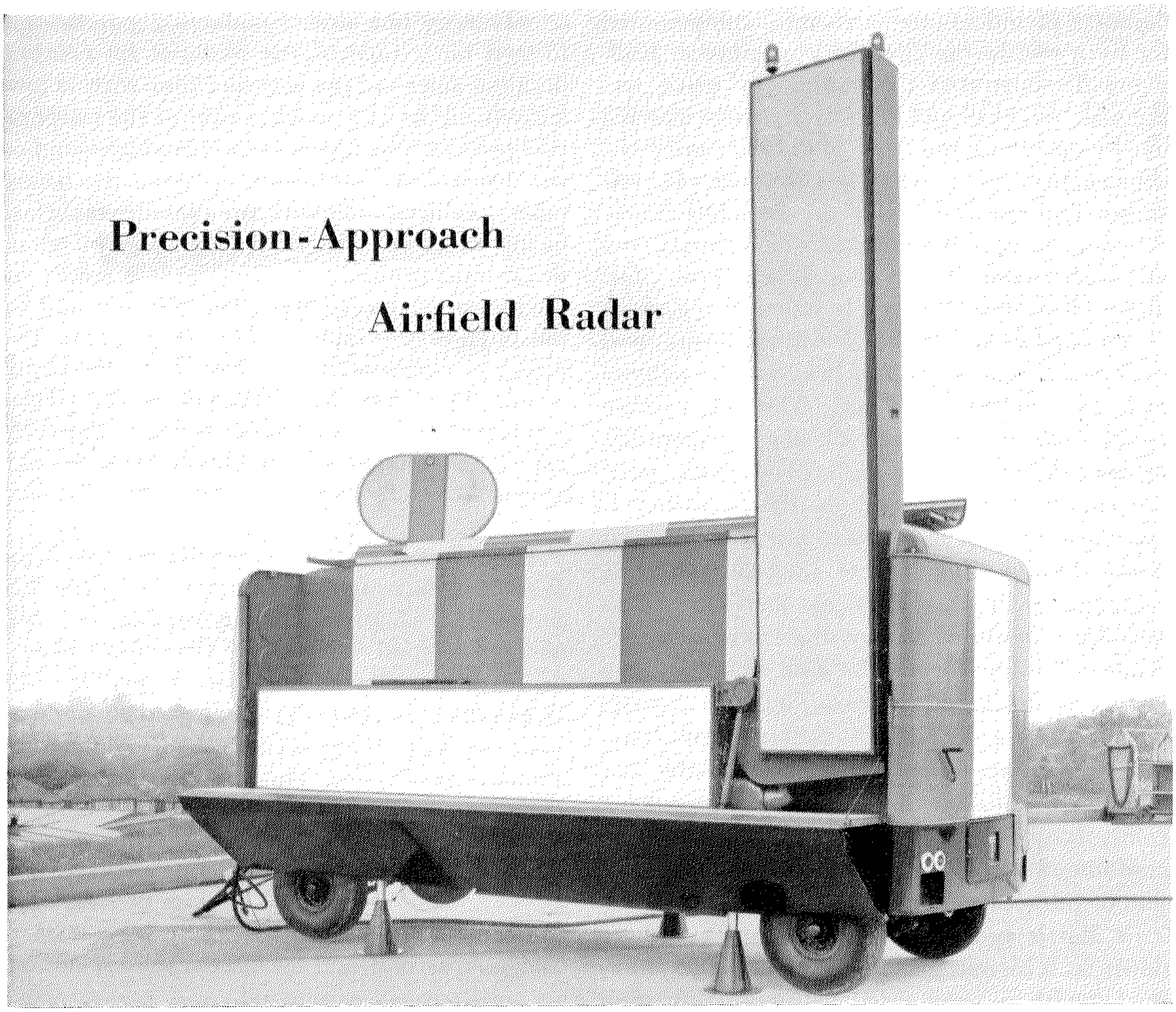
[1] H. E. Kallmann, "Equalized Delay Lines," *Proceedings of the IRE*, volume 34, pages 646–657; September, 1946.

A compensated form of the basic *D* cable was developed utilizing this method of time-delay compensation in a form that is adaptable to continuous factory production. Metallized Mylar film of the type employed for capacitors is used. This film is available in a range of thicknesses down to 0.25 mil (0.00025 inch or 0.006 millimeter) with an evaporated coating of aluminum or silver on one side. Conducting patches are formed on a tape of the material by creating dividing lines on the metallic film with a low-current arc or as a modification of the evaporation process. The tape thus prepared is wound on the delay-cable core before applying the helical inner conductor and with the metallized surface facing the core. The frequency band in which the compensation is effective is determined by the length of the patches along the tape and by the pitch of the tape winding on the core. The degree of compensation is affected by the width of the tape and the thickness of the Mylar insulation. The Mylar tape, even in the 0.25-mil (0.006-millimeter) thickness, has enough strength to allow the use of conventional taping machinery.

The response of a compensated version of the *D* cable is illustrated in Figure 5. This cable has a nominal characteristic impedance of 1900 ohms. The total variation of time delay from 1 to 10 megacycles is less than 3 percent. The attenuation is somewhat higher than for the basic *D* cable, the 3-decibel point occurring at about 5 megacycles. This may be improved to some extent if the metallized Mylar can be obtained with a higher metallic-film conductivity. The effect of the time-delay compensation on the rectangular-pulse response is quite obvious. The rise time of the transmitted pulse is reduced to 0.06 microsecond, and the undesirable oscillations are substantially eliminated.

Precision-Approach Airfield Radar

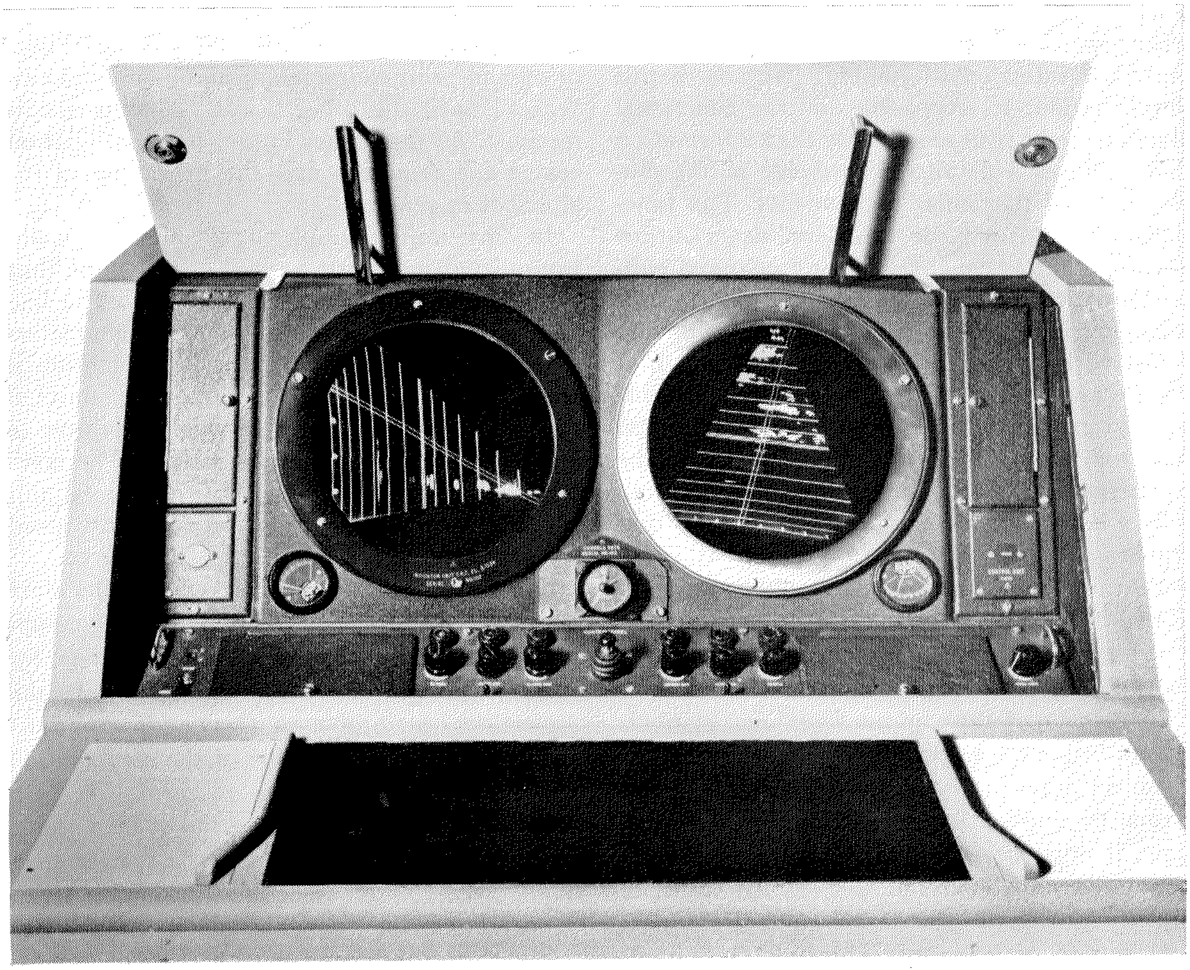


DURING the annual Society of British Aircraft Constructors' exhibition, held at Farnborough, England, the radio division of Standard Telephones and Cables, Limited, presented an operational display of their precision-approach radar equipment. This equipment, probably the most comprehensive of its type in the world, is already in operation at two airfields of the British Ministry of Supply, the International Airport of Zurich, Switzerland, two airfields of the Royal Netherlands Airforce in Holland, and at the Royal Canadian Navy's base at Shearwater, Nova Scotia.

The equipment provides the ground controller at an airfield with visual three-dimensional information about aircraft approaching the runway; distance, azimuth, and elevation are all indicated. This information, vital during inclement weather, enables the controller to convey to the pilot over the normal radiotelephone channel, all instructions necessary to fly the aircraft to the approach end of the runway, whence a safe visual landing can be made.

The radar section of the equipment is installed in an unattended vehicle normally positioned at the side of the runway facing down-wind. Two 50-kilowatt pulse-type radar transmitters operating in the 3-centimeter band radiate fan-shaped beams from two antennas in the direction of the descent path. The 10-mile (16-kilometre) working range of the equipment thus represents a swept space of approximately 3 miles by 1 mile (4.8 by 1.6 kilometres) in which an aircraft can be detected. Radar echoes from the aircraft are interpreted by the receiving equipment in the vehicle and transmitted by a radio link to the airfield control tower, where they are presented to the controller on azimuth/range and elevation/range cathode-ray tubes.

The control tower display console showing the elevation/range cathode-ray tube indicator on the left and the azimuth/range indicator on the right.



Functional-Diagram Approach to Electronics in Telegraphy

By E. P. G. WRIGHT and D. S. RIDLER

Standard Telecommunication Laboratories, Limited; London, England

DESIGNERS of electronic digital computers have provided telecommunication engineers with an attractive method of expression which, on account of its inherent simplicity, is likely to be widely adopted in the future. The basis of this new form of expression comprises a small number of symbols representing clear-cut and frequently used functions. These symbols can be grouped together in different combinations to produce functional diagrams that express the philosophy of the design.

It must be appreciated that such functional diagrams do not attempt to describe the circuit arrangement or even the type of components that it is intended to employ. Because of this simplification, it is possible for the functional diagrams to be read more easily than information that requires a detailed knowledge of the employment of particular components. The functional diagrams may be considered as an outline drawing that is capable of being coloured subsequently to produce a picture. The process of colouring is roughly analogous to filling in the detail with particular components although it will be appreciated that there will be standardized arrangements for many of the functions so that the work of filling in involves chiefly the consideration of interconnecting standard units.

The advantage of the use of functional diagrams is the removal of the detail that is essential for actual operation but unnecessary to describe the general method of operation. It is a striking example of providing a better view of the wood by removing some of the trees.

In this paper the various functional symbols are applied to telegraph systems and the reader is given an opportunity of judging whether these claims to simplicity are justified. At the same time, it must be pointed out that simple functional diagrams will not necessarily translate directly into simple practical circuits by the mere substitution of component blocks. The circuit designer must still use his art although the functional diagram may assist him in the thought process necessary in the evolution of the design.

1. Symbols for Functional Diagrams

1.1 SIMPLE TELEGRAPH SYSTEMS

It will be already understood that in automatic telegraphy it is usual to transmit characters in the form of combinations of 5 units, each unit being either a "mark" or a "space." The 32 combinations available are insufficient to include both alphabet and numerals and for this reason use is made of two combinations known as "figures" and "letters." The reception of one of these combinations causes the receiving teleprinter to print characters from one only of two "cases" until the other "shift" combination is received. As the names imply, one combination covers the numerals and the other covers the alphabet in general.

In start–stop telegraphy the 5 units are usually transmitted serially at a speed of 50 bauds; that is, each unit lasts for 20 milliseconds. The line rests in the mark condition when no characters are being transmitted and each combination is preceded by one additional starting unit of space, which causes the receiving teleprinter to undergo a single cycle of operations. The combination is concluded by another additional unit of mark, which makes allowance for any speed difference between transmitter and receiver.

When a character is being received, the start space must be recognised by the receiving equipment to initiate the cycle of operations for reading the character. During this cycle the theoretical centre of each element is examined to ascertain the constitution of the combination of 5 information elements received. To help visualise an arrangement that might be suitable for carrying out these functions, a trigger such as *1F* and a counter such as *1C* in Figure 1 can be postulated. The trigger is a device having two stable states indicated by the compartments labelled *1* and *0*, which may be set up to the exclusion of each other by signals on the appropriate input leads indicated by arrowheads pointing towards the compartments. The counter, on the other hand, has 140 stable states and is advanced from one state to the next by a signal applied to the single input. The outputs in both cases are indicated by arrowheads pointing away from the compartments. The symbol *1G* denotes a gate, and the encircled *2* indicates that 2 simultaneous inputs are necessary to produce an output.

It is assumed for simplicity that, as the line alternates between mark and space during the course of a character, a signal will appear on an *M* lead when the line is in mark and on the *S* lead when it is in space. In these circumstances, the arrival of the start space will trigger compartment *1* of *1F* to give an output, which by convention will be called *1F1*. *D* is a source of driving pulses always present so that *1F1* causes pulses from *D* to pass through *1G* and to step counter *1C* in response. The counter will continue to step at intervals determined by *D* and to provide outputs timed in relation to the start space until further action is taken.

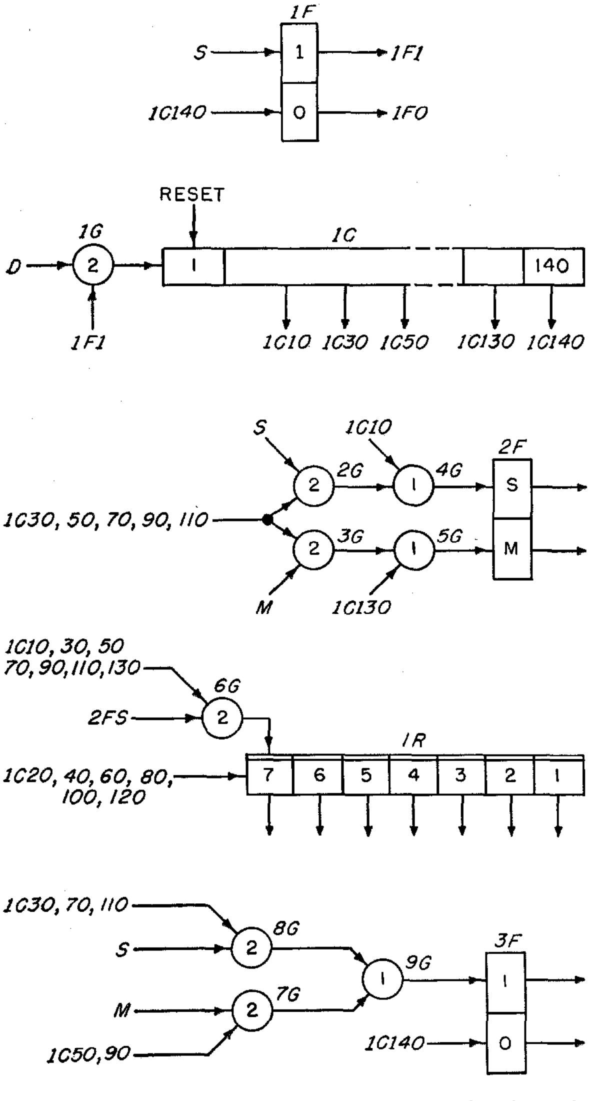


Figure 1—Elements of a start–stop automatic telegraph system. *S* represents a space, *M* is a mark, and *D* is a driving source.

In case the reader may suspect that these symbols express something more than what first meets the eye, it may be noted that a trigger such as *1F* represents the function performed by an electro-magnetic relay with a locking winding arranged so that it may be maintained in either operated or unoperated condition according to some external control. The output *1F1* from the trigger *1F* represents a break contact. The 2-input "and" gate *1G* can be provided by 2 contacts in series, whereas the 2-input "or" gate *4G* described later can be provided by 2 contacts in parallel; the counter *1C* represents the type of function provided by a telephone stepping switch.

With 50-baud transmission, each element in the telegraph character is 20 milliseconds in duration and with the start and stop elements added to the 5 information elements, each character will require a period of 140 milliseconds in all. It is, therefore, desirable that the counter *1C* should take a period of this order to make a complete cycle. The pulses from source *D* are, therefore, made to occur at a rate of 1000 per second, and at the conclusion of the cycle of operations the counter *1C* can be arranged to reset the trigger *1F* from 1 to 0 at 140 milliseconds, thereby disconnecting the drive to the counter, which will cease until the arrival of a subsequent character causes the cycle of operations to be repeated.

It is desirable to inspect the incoming line when the counter is in positions *1C30*, *50*, *70*, *90*, and *110*, as those 5 time positions will represent the theoretical centre of each of the 5 information elements. Inspection at these times will permit the use of a signal of greatest distortion without causing misoperation.

The ability to read a telegraph character may be used for retransmitting or recording a message. It will be appreciated that in the former case it may be desirable to regenerate the signals to correct distortion due to line or atmospheric conditions. In the case of regeneration, a second trigger, *2F*, may be incorporated, having two stable conditions *M* and *S* representing mark and space respectively, the trigger *2F* being altered at the examination times by an input from *2G* or *3G* if it is found that the condition of the line has changed since the last examination. As the start element is always a space and the stop element is always a mark, these two conditions can be applied to *2F* (without reference to the condition of the line) through gates *4G* and *5G* respectively. Furthermore, as the line is to be examined at the centre, as opposed to the beginning, of each elemental period, the trigger *2F* will be changed half an element in time later than the changes are received from the line, and for this reason the start element is applied to the trigger *2F* at *1C10* and the stop element at *1C130*. The gates *2G* and *3G* provide joint control according to the condition of the line and at the proper times of examination, and the output of *2F* represents the mark or space condition to be applied to the outgoing line.

On the supposition that it is desired to read and record the characters, there must be introduced a device that is capable of storing any one of the 32 different combinations that may be produced by the 5 different elements. A suitable device is provided by a stepping register such as *1R*, the symbol for which shows a 7-element register enabling the condition of the start and stop elements to be registered in addition to that of the 5 information elements. The way in which this register is operated is as follows. Each of the register elements is capable of being set in either of 2 conditions, and in particular the element numbered 7 is shown arranged to be set by an input coming in from above. The arrow shown coming in to the side of element *7* represents the stepping means that causes the pattern of the symbol as a whole to make a step forward from left to right so that the condition of element *7* will be advanced to element *6* and so on. Now it will be clear that, if the normal condition of each element represents a mark condition and the operated condition of each element represents a space condition, then the result of the periodic inspection of the line can be applied to element *7* as 7 discrete inputs at the times indicated, the composite pattern representing the character being obtained eventually by providing a step to the register between each examination time. As previously described, the line passing from mark to space starts the counter or time scale *1C*, which can provide examination pulses at *1C10*, *30*, *50*, *70*, *90*, *110*, and *130*, an input being presented to *1R7* at these times if the line is found to be in space. At intermediate times such as *1C20*, *40*, *60*, *80*, *100*, and *120*, a step is provided so that the pattern of marks and spaces gradually extends across the register and at the conclusion of the character its complete identity is available. It will be appreciated that on completion the start element that was initially inserted into *1R7* will have stepped forward to *1R1*, the 5 information elements will be in *1R2*, *3*, *4*, *5*, *6*, and the stop element will be present in *1R7*, where it was introduced but not subsequently moved. The register function has no single mechanical analogy, the function often being performed by a combination of switches and relays.

The four symbols introduced above represent basic functions that can be used in many different combinations to describe a variety of systems not only in telecommunications, but also in industrial applications such as digital computers. To appreciate the possibilities of such systems, it is only necessary to master the

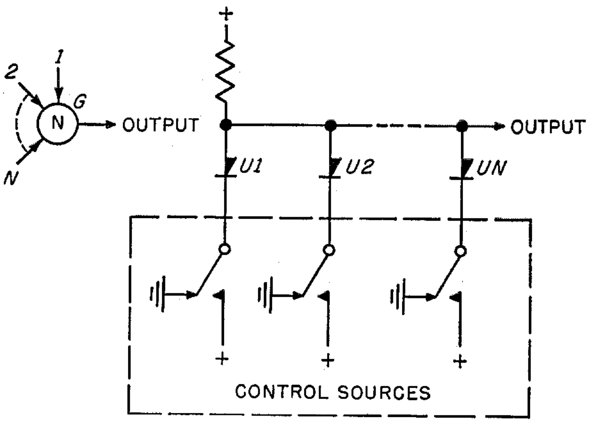


Figure 2—Coincidence (and) gate that produces output only when all control sources apply positive potential to the rectifiers.

simple functions represented by these four symbols. The numbers associated with the symbols may vary to suit the requirements and the components to be used. The number of inputs applied to a gate must depend on the determining factors. The capacity of the counter and register must be chosen to suit their application. Similarly there will be occasions when a trigger device may serve a useful function with more than two elements, provided not more than one is operated at a time.

1.2 Circuit Methods

Having briefly introduced the symbols by way of well-established functions, it is now proposed to describe some of the circuit components and methods by which the symbols may be realized. Necessarily this must be in the nature of a brief review since the advances during the last decade or so have been as numerous as they have been rapid.

It is interesting to note that the computer engineer has made a substantial contribution towards new methods of realizing the functional diagram in much the same way that he has promoted its use. Because the hard valves used in the early digital computers were found to be unsuitable in some respects, it became necessary to seek new methods. The search produced several new solutions that were readily accepted and tried out in this new field.

1.2.1 Gate Circuits and Components

The rectifier, or diode, is perhaps the most commonly used circuit element in gates on account of its inherent simplicity and low cost. A typical coincidence rectifier gate arranged to give an output when a positive potential is applied to all inputs at the same time is shown in Figure 2. The positive voltage from the supply source passes through the rectifiers *U1*, *U2* . . . *UN* to the control sources, shown schematically. If any control is at zero potential, the common connection will be held near zero (the potential drop across the rectifiers). On the other hand, if all sources are simultaneously positive, the common connection will become positive to indicate coincidence. The corresponding mixing (or) circuit is shown in Figure 3 in which any single positive control source will cause a positive output across the resistance. These circuits might make use of hard valves or metallic rectifiers such as selenium or germanium for example according to the design requirements in respect of cost, reliability, speed, et cetera.

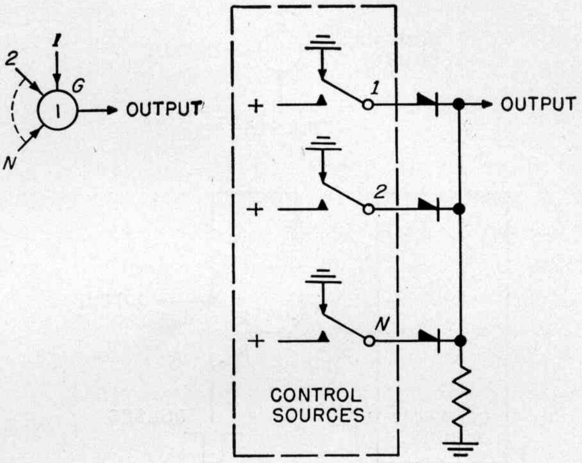


Figure 3—Mixing (or) gate that produces output if any one control source applies a positive potential to the rectifiers.

It is characteristic of the rectifier gate that it produces an overall power loss. Often this does not inconvenience the designer since the trigger-circuit components may contain an inherent gain, or if this is not so a suitable pulse regenerator may be inserted. Nevertheless, a useful class of gating circuit is that in which a supply of pulses can be switched from a low-impedance source without appreciable loss while the control source uses only a minimum of power.

For example, a gate circuit might comprise a hard valve with a cathode-connected source of negative pulses that can be "switched" to the anode by means of a suitable control voltage on the grid. Alternatively, a cold-cathode gas trigger tube can be used in much the same way. A third example using a more-recently developed component is shown in Figure 4 in which use is made of the switching properties of a point-contact transistor.[1] Negative pulses from a low-impedance source are applied to the collector and may be switched to a load *R1* in the emitter under the control of a potential applied to the base resistance *R2*. If this potential is positive, no current will flow into the emitter, and in consequence the collector current, when the pulse is on, will reach only a small value. On the other hand, if the control potential is negative and the current gain sufficient, then, when a pulse is present a small initial current in the emitter will

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[1] British Patent Application 29848/53

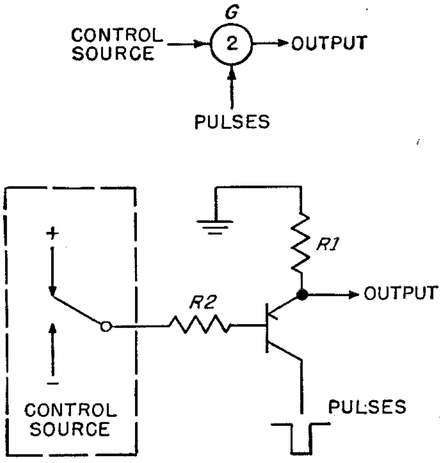

Figure 4—Arrangement of a point-contact transistor to produce a coincidence (and) gate.

cause a larger current to flow in the collector, thus further reducing the base potential until the transistor "bottoms" and is held on by the difference between the emitter and collector currents flowing in *R1*. In these circumstances an effective low impedance will be maintained between collector and emitter and the pulse current will be determined mainly by the load.

Junction transistors can also be used for this class of gate although the technique of using them is rather different from that outlined for the point-contact type.

Mention should be made of the magnetic gate, which may have considerable importance in the future on account of its ruggedness, speed of operation, and low cost. Use is made of the properties of the rectangular hysteresis loop shown typically in Figure 5. If a suitable constant-current pulse source is applied to one input winding of a toroid such that the hysteresis loop is only traversed between point *1* and point *A*, then only a very small electromotive force will be induced in the output winding since there can be very little change of flux. On the other hand, if the control winding is energized with a similar value of ampere-turns at the same time as a pulse occurs, then the hysteresis loop will be traversed to point *B* and an output pulse will be induced in the output winding. In this way a single pulse has been gated through. Steps must be taken to restore the magnetic state of the core from state *0* to state *1* in order that the cycle may be repeated. This can conveniently be achieved by interposing pulses of opposite polarity between those from the pulse source.

In practice, magnetic gates can use ferrite toroids producing rectangular hysteresis loops, which permit operation at very high speeds. Furthermore, the windings are often formed by only single turns so that the construction is very simple.

1.2.2 *Trigger Circuits and Components*

In the last decade so many solutions have been proposed for what may be called the trigger group that only a few of the more-promising components need be discussed in any detail.

Hard-valve trigger circuits, such as the Eccles–Jordan, are well known and extensively used and probably represent the best of current practice. However, there is much interest in replacing them for reasons of power and heat dissipation, reliability, life, initial cost, and general convenience.

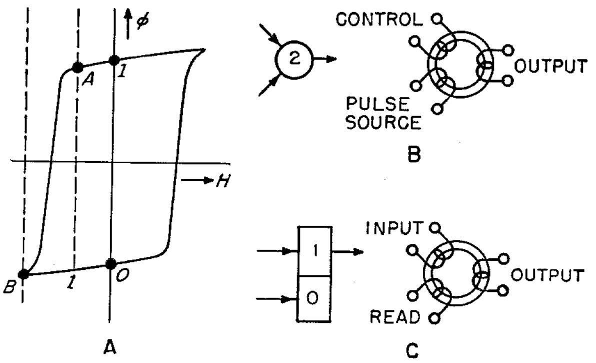

Figure 5—Typical hysteresis loop *A* on which the magnetic versions of the gate *B* and trigger *C* are based.

For comparatively slow-speed circuits the gas tube is attractive since the properties of the gas discharge are well adapted to a 2-state device and since the visibility of discharge can be a valuable aid to maintenance. The development of gas-tube circuits has also provided some experience with useful techniques that can similarly be applied to some transistor circuits that will be outlined later.

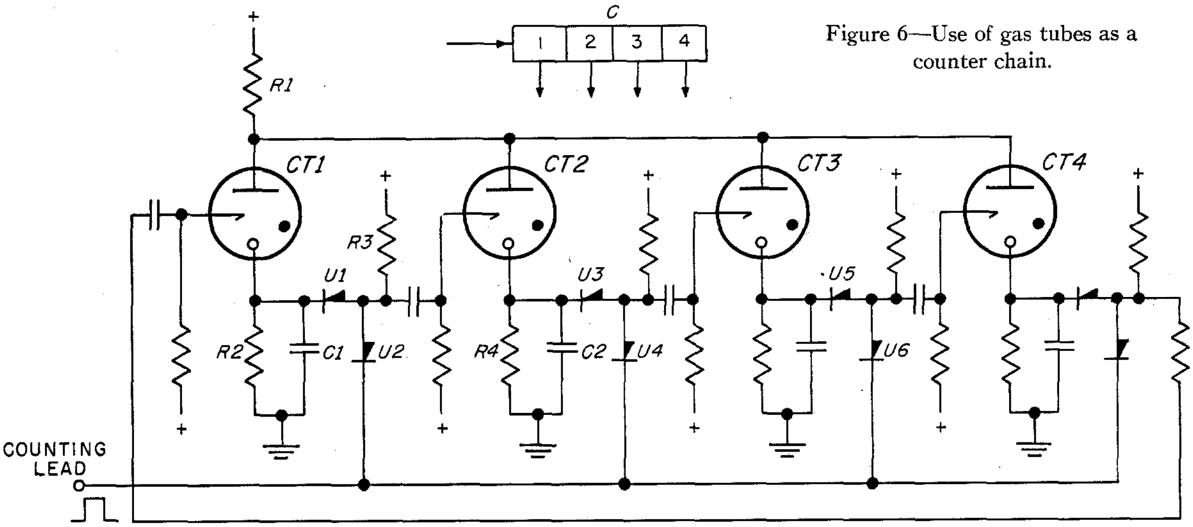

Figure 6—Use of gas tubes as a counter chain.

Since a discharge in a gas can be extinguished only by disconnecting or reducing the anode voltage, it has become common to use tubes in pairs or groups, the firing of one gap being used to extinguish another. This is not, however, an inconvenience, since two outputs corresponding to *1F1* and *1F0* of Figure 1 are thereby provided. When one tube of the pair is rendered conducting by the application of a triggering potential, the instantaneous flow of current can be used to depress the voltage across the other and thereby extinguish it.

The process can be made clearer by reference to a rather-more-complex circuit. Figure 6 shows 4 cold-cathode trigger tubes *CT1–4* arranged as a ring counter. Assume that one tube, say *CT1*, is conducting and that current is flowing from the positive supply through the common anode load resistor *R1*, the main gap of *CT1*, and resistor *R2* to earth. Capacitor *C1* will have charged to the potential across *R2*, which also determines in conjunction with *R1* the anode voltage supply to the other tubes. If, now, a pulse is applied to the common counting lead, then only the rectifier coincidence gate formed by *U1*, *U2*, and *R3* produces an output to trigger tube *CT2* since only the positive plate of *U1* is positive due to the drop across *R2*. The resultant current passing through the main gap of *CT2* depresses the voltage on the common anode connection due to the presence of *R1* and, since *C1* was charged and *C2* not, the potential drop across *CT1* is reduced below the glow-maintaining value. With proper choice of constants, tube *CT1* will extinguish while *C2* charges with a time constant:

$$\frac{R1 \cdot R4 \cdot C2}{R1 + R4}.$$

Meanwhile, *C1* discharges through *R2*. When these charge and discharge processes are complete, the glow discharge may be further advanced from *CT2* to *CT3* by the application of a second pulse and so on.

A more-compact and convenient counter can be made by the inclusion of the gas gaps within a single envelope[2] if use is made of directional ionization coupling among, say, 10 main cathodes and 10 commoned transfer cathodes. Figure 7 illustrates a typical arrangement for 4 main cathodes.

Each cathode has geometrical asymmetry, so that, for instance, if current is initially flowing through the common resistor *R1*, gap *AK3*, and resistor *R3* to earth, only the transfer-cathode gap *AK4* will have a low breakdown potential due to the proximity of the discharge. Therefore, if a negative pulse is applied to the transfer cathodes, only gap *AK4* will break down and gap *AK3* will extinguish, due to the additional voltage drop across *R1*. Now, since the gap *AK5* will come under the influence of the glow on *AK4* while gap *AK3* deionizes, when the pulse goes off, gap *AK5* will break down and conduct.

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[2] G. H. Hough and D. S. Ridler, "Multicathode Gas-Tube Counters," *Electrical Communication*, volume 27, pages 214–226; September, 1950.

In this way a pulse applied to the commoned transfer electrodes will have caused the discharge to take a step forward.

Gas-tube trigger circuits may be considered practical up to a speed range in the region of 10 000 steps per second although higher speeds have been attained for stepping registers. The input and output characteristics of the circuits are convenient, particularly for rectifier gates. For a further increase in speed, the transistor is probably the most convenient and versatile element.

The point-contact transistor is a most-convenient type for trigger circuits since it has the property of being able to hold itself conducting in a suitable circuit and can, therefore, conveniently fit one compartment of the functional symbol. However, it is apparent that there is a considerable field for the junction type, which is less sensitive to temperature than the point-contact unit and, therefore, more likely to find favour where the ambient range is wide. The issue is by no means decided, and it is probable that there will be an extensive field of application for both types and perhaps for some new varieties. Whereas the circuits of junction types have much in common with those for hard valves, the approach to the use of the point-contact transistor is rather novel and will be described in view of its greater interest.

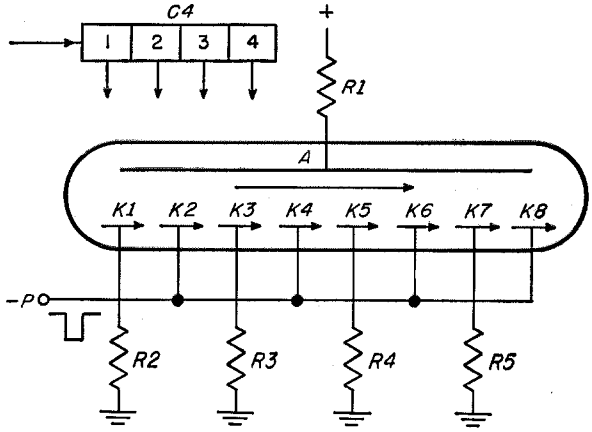

Figure 7—In this counter, several sets of gas gaps are mounted in a single envelope.

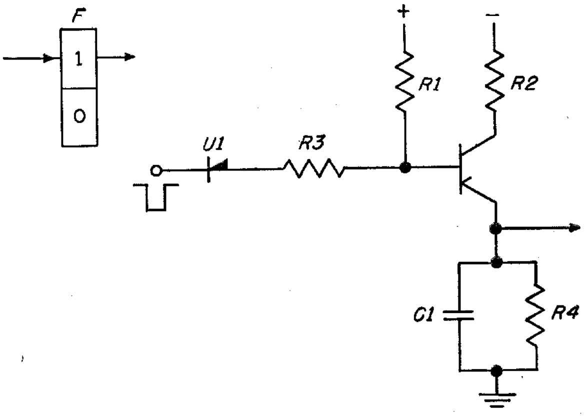

Figure 8—Trigger circuit using a point-contact transistor.

The most suitable point-contact transistors for trigger functions should have a current gain between the emitter and the collector at least greater than unity and should have a low standing current in the collector with zero or a small negative voltage on the emitter. Given these conditions, satisfactory trigger circuits can be evolved. Figure 8 shows one simple example of such a circuit, which is somewhat analogous to the gas-tube trigger and has similar desirable properties from the point of view of ease of interconnection.[3] In the "off" condition, the current flowing through *R1*, *R3*, and *U1* is arranged to be somewhat larger than the small collector-to-base current and of opposite polarity so that the base is slightly positive and the collector, therefore, presents a high impedance. If, now, a negative triggering pulse is applied such that emitter current flows, then current will also flow in the collector and the circuit will take up a second stable state in which the transistor is held on by the difference in current between emitter and collector that maintains the base negative. The circuit can be triggered back to the off condition by momentarily reducing the collector supply to zero, which can be conveniently achieved by means of a second transistor trigger circuit.

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[3] British Patent Application 33618/53.

In much the same way that the output from the cathode circuit of a gas tube may be used to control the triggering of a further tube as shown in Figure 6, so the output from *R4* in parallel with *C1* in the emitter of the transistor in Figure 8 may be used to control the triggering of a further transistor via the base. Again this may be illustrated by considering the transistor ring counter in Figure 9.

Assume that one particular transistor such as *T1* has been triggered into the "on" condition so that the current flowing through it will result in a negative potential across *R1*. If, now, a stepping pulse is applied to the common counting lead, then only the coincidence gate formed by the rectifiers *U1* and *U9* will produce a negative output pulse since the other gates will be held at earth through the emitter loads. Transistor *T2* will, therefore, trigger and the additional current necessary to charge *C2* will cause the potential across *L* to rise and restore *T1* to the off condition.

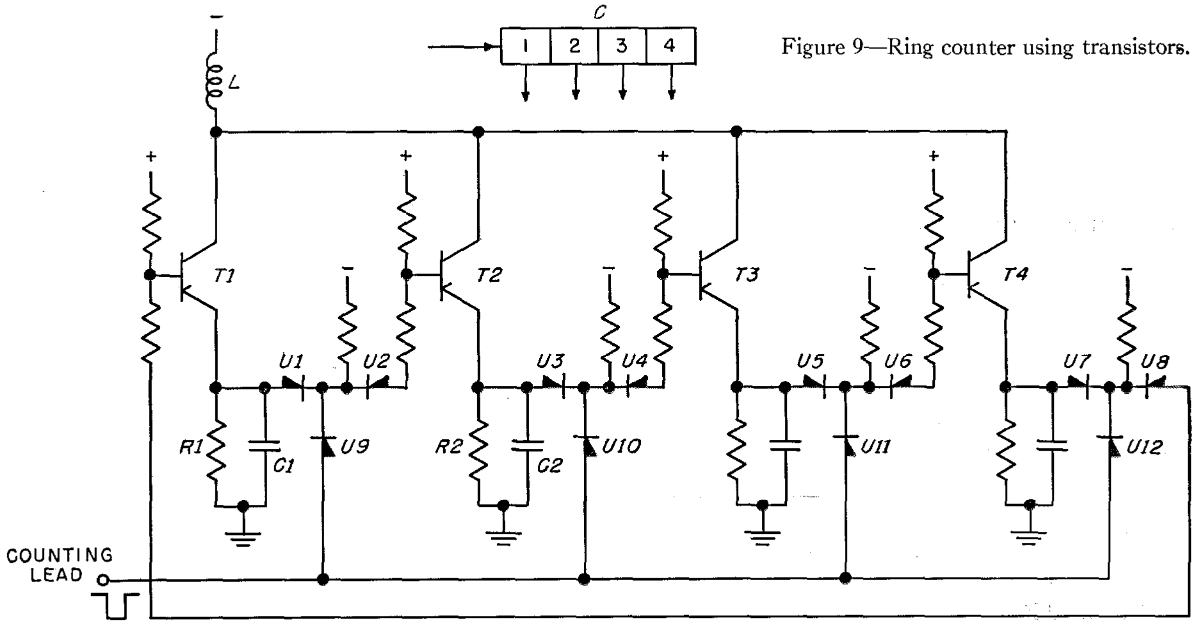

Figure 9—Ring counter using transistors.

It is desirable to use a common inductance rather than a common resistance, as in the case of the gas-tube counter, since the variation in the off-condition collector current between transistor samples would give rise to an unpredictable collector supply voltage on account of the potential drop across the common resistance.

The design of these basic gas-tube and transistor circuits, which are in many respects similar, has been influenced by the functional diagram in the sense that they are flexible elements easily interconnected by virture of their input and output characteristics. At the same time it is necessary to regard the functional diagram itself as flexible and capable of alternative interpretations. For instance, the counter *1C* of Figure 1 might in practice be broken down into 3 stages, the count of 140 milliseconds being derived by counting pulses occurring at a rate of 5000 per second, firstly in tens to derive 2-millisecond periods, then this output in tens again to obtain 20-millisecond periods, and finally in sevens to obtain 140 milliseconds. The interconnection of the three counters necessary for this process is simply achieved by means of gates, and any time interval may be obtained by outputs in combination. Even so, the most economical circuit may not result when the functional elements are directly replaced by circuit elements, and it is here that the circuit designer must use skill in the arrangement of components.

In addition to the types of trigger already described, the static magnetic trigger, like the magnetic gate, is an element of growing importance, although circuits using this device differ in some respects from circuits using gas tubes and transistors and may not fit so easily into the functional diagrams outlined.

Magnetic triggers make use of the two states of remanent magnetism denoted *1* and *0* on the hysteresis loop shown in Figure 5. The main feature of this loop is that the remanent flux is very nearly the same as the saturation flux. If a suitable pulse is applied to the input winding of a core made from material with this type of hysteresis loop, then it will be left in either state *1* or *0*, dependent on the polarity of the pulse. There is no standing indication of the remanent state, and to provide an output, it is necessary to pulse the core a second time and to observe the output winding for the presence of an electromotive force indicating a change of flux. If this second, or reading, pulse has the same polarity as the original input pulse, then the flux change will be small and only a small electromotive force will be induced across the output, or reading, winding, whereas if the original pulse were of opposite polarity the hysteresis loop would be traversed towards the other remanent state and a resultant output pulse would be induced in the reading winding. A simple trigger built on this basis has the disadvantages that there is no standing output and that any interrogation results in a return to a particular state. On the other hand it has the advantage of no standing power drain.

Research is now being directed towards methods of reading the polarity of the remanence without destroying it, and there would appear to be reasonable hope of success. If these methods prove simple, then there can be little doubt that magnetic elements will be very versatile, very rugged, and have a long life.

2. *Functional Diagrams Applied to Telegraph Systems*

Some of the less-well-known operations in telegraphy will now be described by means of functional diagrams that will use the symbols outlined in section 1.

2.1 CHARACTER IDENTIFICATION

For supervisory purposes a message in process of transmission may have to be scanned for a particular character, which may, for example, represent the end of the message. It would clearly be possible to record each character in turn on a register such as *1R* in Figure 1 and when the time scale has given the sixth step to *1R* to examine the significant positions for the particular combination that requires special attention. However, this function can be performed much more simply in the following manner.

It is possible with a trigger such as *3F* to indicate with certainty if a character received is *not* the one requiring special attention; thus, if an indication from *3F* is not given, it follows that it was in fact the special character. The functional operation may be described in more detail as follows. Supposing that the particular character to be observed is *M S M S M*, then if the first element of the character is a space it is certain that it is not part of the special character and a transfer can take place from *3F0* to *3F1*. Similarly, if the second element is a mark, which cannot possibly be a part of the special character, then the trigger *3F* can be transferred from *3F0* to *3F1*. In a similar manner the third and fifth elements can be treated as the first element, and

Characters Transmitted—

SMSMSMM *MMMM* **SMSMMSM** *MMM* **SMMMMSM** *MMMMMM* **SMMSSSM**

Additional False Start—

SMSMSMM *MM* $\underline{\text{S}}$ *M* **SMSMMSM** *MMM* **SMMMMSM** *MMMMMM* **SMMSSSM**

Characters Received—

SMSMSMM *MM* **SMSMSMMSMMMMSM** *MMM* **SMMMMMM** *M* **SMMSSSM**

Characters Interpreted Wrongly Due to Single False Start

Figure 10—Effect of a false start element introduced by interference. Each 7-element character shown in boldface starts with a space S and ends with a mark M. The spurious start signal is underscored and causes the reception of three wrong letters in the place of the two transmitted characters.

the fourth element can be treated in the same way as the second element. If, after the last examination at *1C110*, there remains an output from *3F0*, then the special character *M S M S M* has just been received.

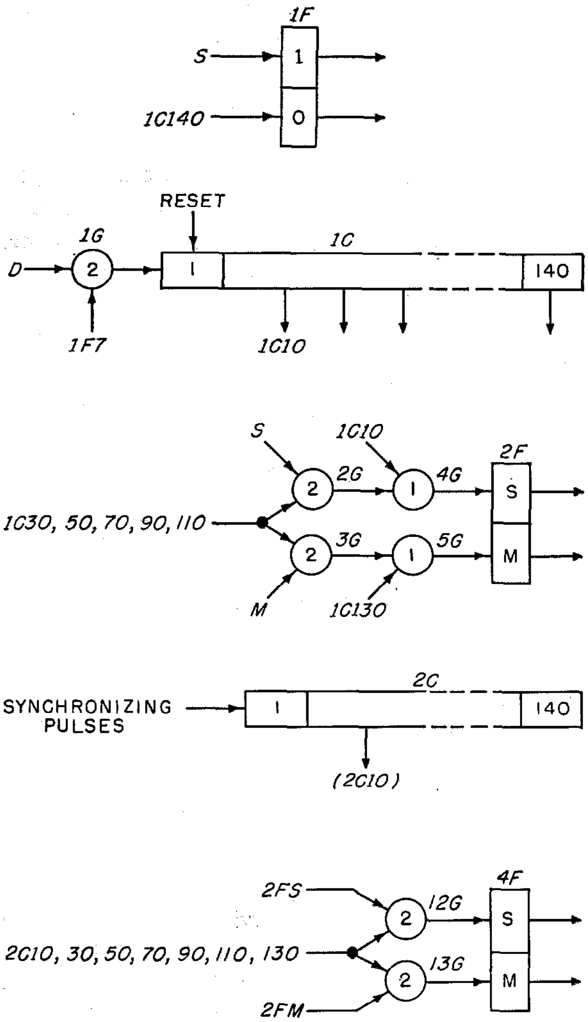

Figure 11—Transmitting end for element synchronization.

2.2 Synchronous Telegraph Systems

It is well known that in start–stop telegraph systems one of the most-damaging effects of interference is to cause the receiver to become out of phase with the transmitter, with the consequence that the transmitted characters are misinterpreted until correct phase is regained (see Figure 10). One satisfactory method of avoiding the false start signals, which are responsible for this trouble, is to make use of synchronous transmission, that is to run the receiver in phase with the transmitter independently of the start–stop signals. With such an arrangement it is possible to disregard apparent starts that occur at improper times. Either element synchronization or character synchronization can be used. Element synchronization is advantageous on account of its ability to reduce loss of synchronism, but character synchronization altogether eliminates such trouble although the equipment needed is somewhat greater than that required for element synchronization, as illustrated below.

For element synchronization it is necessary to include triggers such as *1F* and *2F* and a counter such as *1C* (Figure 1) for the recording in turn of each element of the start–stop character. A second counter such as *2C* in Figure 11 is employed to give the synchronous time, and an extra trigger *4F* can be used to transfer the condition from *2F* under control of the synchronous time scale. In this way each element of each character is re-aligned to fit the synchronous times. At the incoming end (Figure 12) a synchronous time scale allows the trigger *5F* to

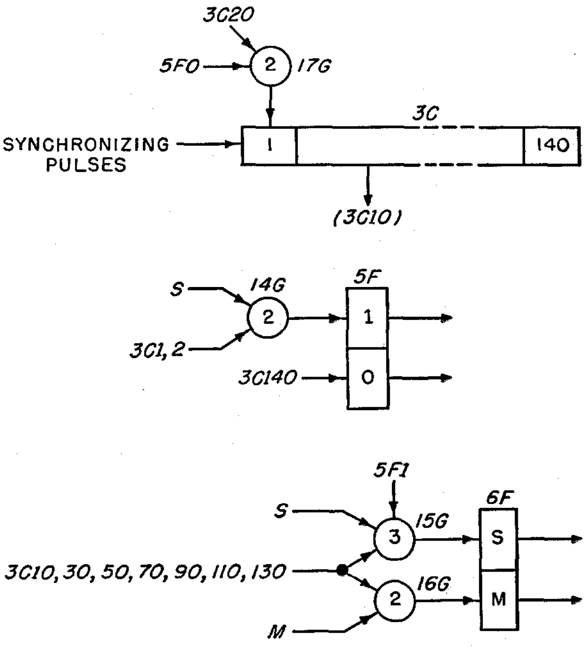

Figure 12—Receiving end for element synchronization.

change over from *0* to *1* only if space is being received at the acceptable inspection time. If there remains an output from *5F* after this period ceases, the time scale can be reset at the proper time to maintain synchronism in preparation for the receipt of a character possibly commencing in the next element position.

With character synchronization it is necessary to store the whole character at the outgoing end to provide for an adverse relation between the start–stop and synchronous time scales. This can be achieved by storing the element values on 7 separate triggers, *11F* through *17F* of Figure 13, which are operated under control of the start–stop time scale *1C* and subsequently read by the synchronous time scale *2C*. It is evident that these triggers must be restored to normal after they are read so that they may be available without delay for re-use by a subsequent character. At the incoming end (Figure 12) no time-scale-reset requirement arises as the trigger *5F* can only pass from *0* to *1* at the start of each cycle of *3C*.

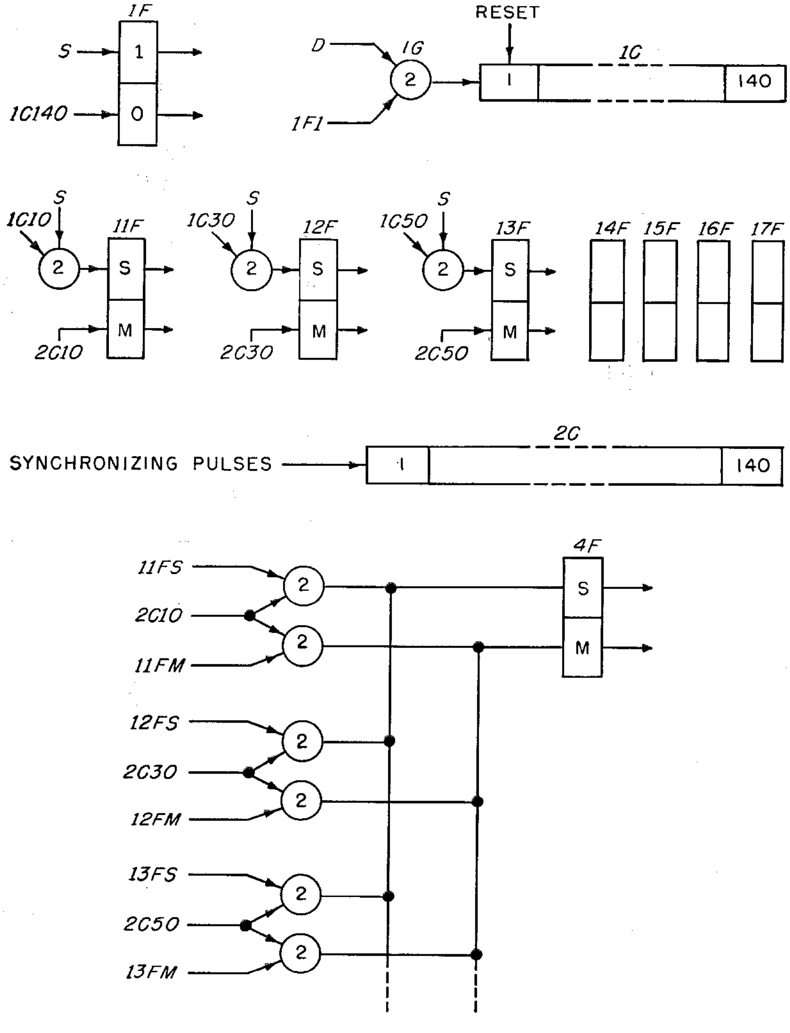

Figure 13—Receiving end for character synchronization.

2.3 Error-Indicating Systems

It will be appreciated that character synchronization virtually eliminates message distortion due to false start elements. However, disturbances on radio, and even line, circuits may occur and result in an incorrect character being received. Various systems have been devised to indicate if a character received contains an incorrect element. The simplest process introduces a parity check; such a check is achieved by ensuring that each character contains an even number of marks or spaces. Suitable equipment can be employed to indicate any character received that fails to satisfy the parity check. The following method may be used to ensure that all characters have an even number of marks.

It is evident that, of the 32 different combinations provided by 5 information elements, 16 of the 32 will contain an odd number of marks, but if the characters formed from these 16 combinations use mark for the stop element, these characters will then have an even number of marks. The other 16 combinations can use a space for the stop element, thereby retaining also an even number of marks. The polarity of the stop element is not important, provided there is always a transition of polarity for the start element of the following character. It will be appreciated that a 2-position counter can determine the appropriate stop and start elements so that each character can be made to contain an even number of mark elements and to ensure that each start element differs from the previous stop element.

For character-synchronous systems in which it is necessary to store each character, the checking principle can be further developed using both start and stop elements for checking purposes. One favoured system is to transmit 32 different combinations of 7 units, each comprising 3 spaces and 4 marks or vice versa. Each character can be checked for the total of marks and spaces and, unless there has been a transposition of a mark element to a space and also a space to a mark in the same character, any normal fault can be indicated. Such systems need to make a character translation from the 5-unit to the 7-unit base and a further re-translation at the receiving end. As a consequence, two relatively involved terminal equipments are needed.

An alternative, and simpler, arrangement is achieved by scanning the 5-unit characters in turn to avoid translating all those characters that contain 1, 2, or 3 spaces and to obtain the constant total by the addition of the appropriate combination of balancing elements to replace the start and stop elements. The seven 5-unit character combinations containing 0, 4, or 5 spaces cannot be treated in this way and must be translated.

2.4 Error-Correcting Systems

Another arrangement aims to determine an element received incorrectly and to correct it so that there is no need to use a channel in the reverse direction to seek a repetition.[4,5] As in the error-indicating systems, it is accepted that certain error combinations will remain undetected, but it is confirmed by statistics that a large percentage of all errors are isolated and could, therefore, be corrected automatically. This alternative is described hereafter as an exercise in the use of functional diagrams to describe a system.

The system considers each pair of characters as a unit for which there are 4 check elements (2 start and 2 stop elements) available for check and correction. These 4 elements provide 16 different combinations that, being transmitted with a pair of characters, describe their constitution on a parity basis.

Assume that the 5 information elements of the first of a pair of characters are lettered *11R1* . . . *11R5*, that those of the second of the pair are lettered *12R1* . . . *12R5*, and that the check elements are numbered *21C*, *22C*, *23C*, and *24C*. A space occurring in the information elements is used to control the check elements as shown in Table 1.

TABLE 1

Control of Check Elements

Space In	Provides Impulses To
11R1	*21C, 22C*
11R2	*21C, 23C*
11R3	*21C, 24C*
11R4	*21C, 22C, 23C*
11R5	*21C, 22C, 24C*
12R1	*21C, 23C, 24C*
12R2	*22C, 23C*
12R3	*22C, 24C*
12R4	*22C, 23C, 24C*
12R5	*23C, 24C*

The information elements of a pair of characters are examined in transmission, and 4 check elements are prepared according to the above list. If, now, a check element (*21C*, *22C*, *23C*, or *24C*) receives an odd number of impulses, a mark is transmitted; otherwise a space is transmitted. If the information elements are similarly examined on reception and check elements again prepared, then the two groups of check elements may be compared to indicate and correct a possible fault. Since an impulse is sent to two or three check elements when a space is present, then if one, and only one, check element compares incorrectly it can be assumed that it is the check element that is wrong, and no correction is necessary. Any one of the 6 different combinations of 2 check elements wrong or of the 4 different combinations of 3 check elements wrong indicates that a particular information element needs correction in accordance with the list above, whereas all check elements wrong indicates several errors that are beyond correction.

Turning now to the functional diagram of Figure 14 and assuming *11F* to be the trigger, *11C* the start–stop time scale, and *12C* the

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[4] British Patent 710258; September 22, 1954.

[5] British Patent Application 34415/53.

synchronous time scale as before, *11R* and *12R* comprise a couple of character registers introduced alternately by the 2-element counter *13C*, which is driven once as each character arrives. The registers *11R* and *12R* record the space elements encountered under control of the start–stop time scale *11C* as described before. In Figure 15, *31F* provides a sequence control and is in form similar to the trigger *11F* except that there are 6 positions, only 1 of which may be operated at a time. Normally in position *31F1*, it advances to *31F2* when the second of a pair of characters is in process of being received. It then awaits the end of a synchronous cycle when it advances to *31F3* in which the 5 information elements in *11R* are emitted; thereafter it passes into *31F4* for 2 check elements. *31F5* and *31F6* control the emission of the second of the pair of characters in a similar way, after which *31F* advances to *31F1* to await the time when a further pair of characters can be transmitted.

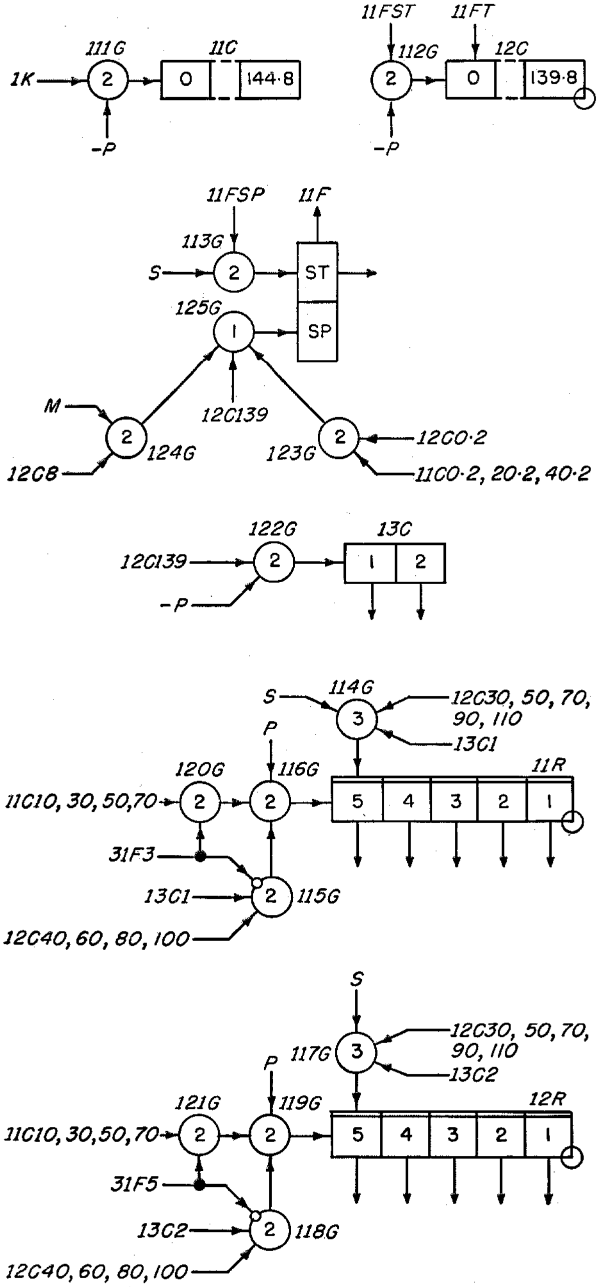

Figure 14—Error-correcting-system character reception, outgoing terminal.

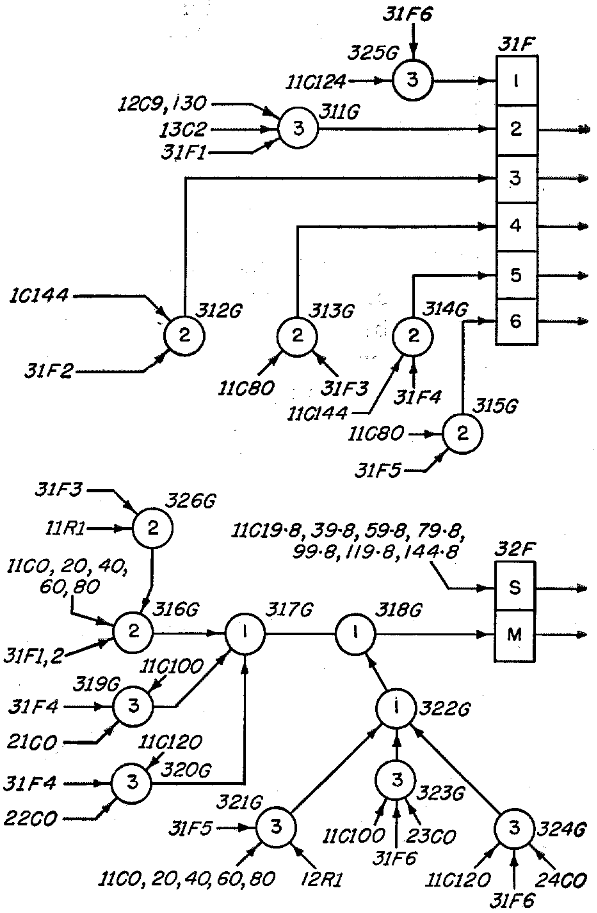

Figure 15—Error-correcting-system character emission, outgoing terminal.

The counters *21C*, *22C*, *23C*, and *24C* of Figure 16 are operated in accordance with the arrangement described above during the time cycle of *12C* (Figure 14) while *31F* (Figure 15) is in position *3* or *4*, that is while the first of a pair of characters is in process of transmission.

The trigger *32F* performs the function of *2F* in earlier examples; it reads the space condition of the registers *11R* and *12R* and of the counters *21C* through *24C* under the sequence control *31F* and the time scale *12C*. *32F* is shown biassed to mark for each element but is changed to space as required.

At the receiving end, the synchronous time scale *41C* of Figure 17 and the registers *41R* and *42R* for a pair of characters follow the technique already explained; the changeover between these two registers is controlled by *42C* which only passes to *42C3* when the trigger *41F* confirms that a character has been received in *41R*. The check counters *43C–46C* are pulsed from *41R* and *42R* during the cycle when the second character is being received. The check elements are also counted, and, if no failures have taken place, each counter will be in its *1* condition: the counter *51C* (Figure 18) will in consequence transmit on the two following cycles the exact contents of *41R1–5* and *42R1–5* through *502G*, et cetera, the standard start and stop elements being inserted. If any combination of 2 or 3 of the counters is in its *0* condition, the appropriate element will be corrected by a pulse applied to *51C*. The trigger *51F* is used to indicate the total number of incorrect counters and to ensure through *506G–509G* that the only correction effected is determined by the precise arrangement of all the counters *43C–46C* that register disagreement through *510G–519G*. If all 4 counters are wrong, there is no correction but the condition can be alarmed and the doubtful character indicated. It should be explained that the registers *11R* and *12R* (Figure 14) are stepped both from the start–stop time scale *12C* and from the synchronous time scale *11C*, the control being vested in *31F3* and *31F5*. The circles surrounding these inputs to *115G* and *118G* indicate that the operation of these gates is inhibited by and not associated with these inputs. It is undesirable that the gates *114G* and *120G* should conduct simultaneously, and such an event is avoided by *123G*, which prevents the unwanted phase relation between *11C* and *12C*. This correcting system needs no translation or message storage before transmission.

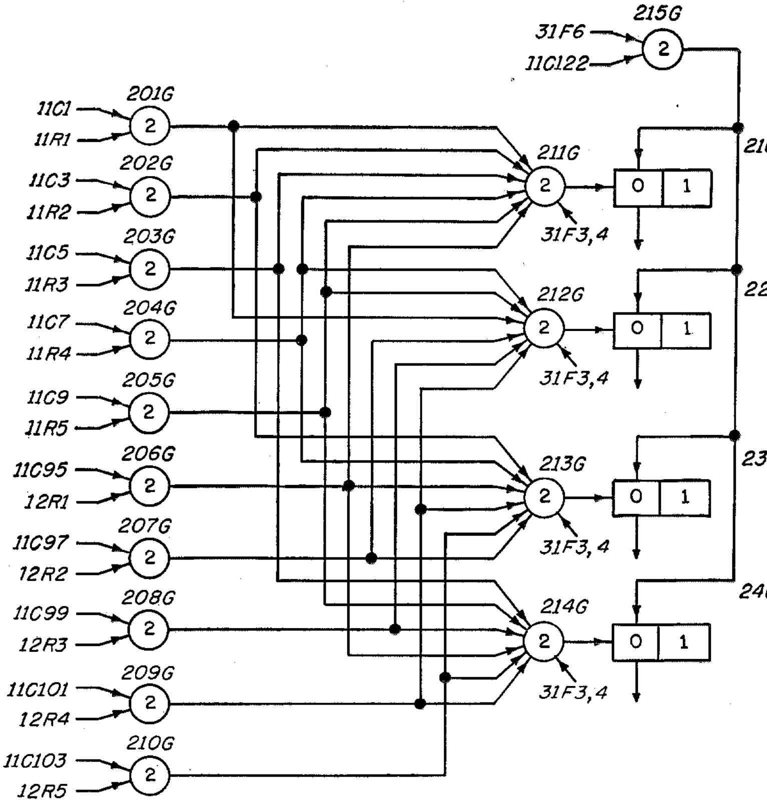

Figure 16—Error-correcting-system check-element determination.

2.5 Extension of Uniperiod or Element Length

It will be appreciated that another way of using the 140 milliseconds of available time in a synchronous system is to extend each of the 5 elemental periods to 28 milliseconds. The characters can be stored in pairs, as described, and the synchronous time scale used to examine the stored values every 28 milliseconds. The longer elements would allow, say, three examinations at the receiving end, and the majority condition would be accepted.

3. *Miscellaneous Applications*

3.1 Flexible Time Scale

A flexible time-scale arrangement is useful for several requirements in telegraph operation. For example, a false-start condition is unlikely to persist for 20 milliseconds, and if it is found to have ceased after a few milliseconds the time scale can be reset and the false character rejected. For passing supervisory signals over a telegraph channel, it is common practice to allow the line to rest in space instead of mark. If then the time scale is arranged to examine the 7 elements, it can ensure that the space condition is maintained until mark reappears before allowing the line to revert to normal operation. When no such signals are in use, it may be desirable to insert the mark for the stop element, disregarding the condition of the incoming line.

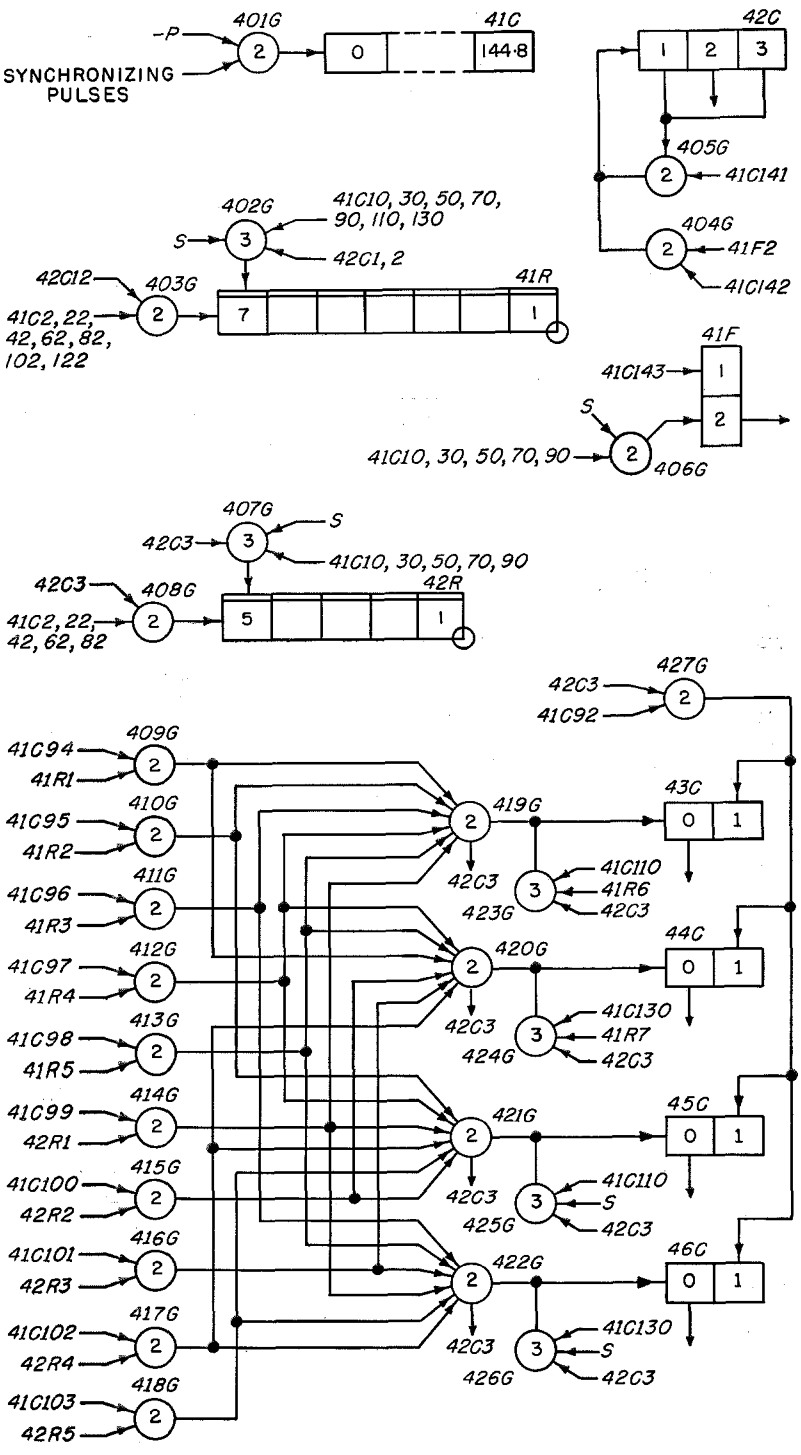

Figure 17—Error-correcting-system character reception, incoming terminal.

Many of these time-scale conditions can be easily varied for experimental and statistical examination to ascertain the most suitable values in different line conditions. For exact comparison work, it has been found essential to make high-quality magnetic recordings of typical distorted messages that can be re-read without further distortion due either to random or systematic variations in tape-reading speed. Tapes can be perforated to determine the different parameter values by the use of special measuring equipment.

3.2 Multiple Regeneration

Although originally introduced to avoid excessive distortion on long telegraph connections, it seems likely that there will be some application for regenerators on quite short lines provided that, as a consequence, the terminal machines can be guaranteed improved operating conditions.

Such improvement would allow the machines more margin, with a consequential decrease in maintenance. However, for such a policy to be attractive it would be necessary to find a very cheap and reliable regenerator with low maintenance expense. One attractive method of meeting these requirements is a form of multiple regenerator in which 20, 40, or 60 circuits could share not only a common power plant but also other common facilities.[6]

For example, with a memory type of store, in which the elements are examined periodically and in sequence, a group of time scales could be built in which the individual parts might be of extremely simple construction. In principle each scanning of the memory could be made to accumulate a counting operation that had been initiated by a trigger such as *1F*. If the repetition rate of scanning is high, the relative position of scanning when the trigger operates represents an error that can be neglected. With a repetition frequency of 0.625 millisecond, a total of 32 scans would mean 20 milliseconds, and 7 periods of 20 milliseconds would cover the normal time cycle. Counting in binary notation, the whole scale could be provided by 8 elements, a number of circuits could be provided on a single memory device and the individual portion of the circuit would be limited to triggers such as *1F* and *2F*.

3.3 MESSAGE STORING AND RETRANSMISSION

The memory type of store may be conveniently utilized for the storage of messages and take the place of a reperforator. In a similar way the transmission of a telegraph message stored electronically involves only the reading of the stored information and means to control the signal sent to line as already described. The scanning speed will normally be far greater than that essential for telegraph purposes, and as a consequence an annex system may be introduced to provide a ready means of altering the time scale. It will be evident that such an arrangement provides facilities for the transmission of messages at different speeds to suit the different conditions that arise. Associated with the need to store telegraph messages, there will naturally be the need to number the messages and to handle them sequentially. Both these functions can be adequately carried out by the memory type of storage using the facility for rapid scanning to achieve the rapid selection of number or message that is naturally desirable for efficient operation.[7]

4. Memories

Mention has been made of the use of a memory type of storage that could be provided by mercury delay lines, magneto-strictive lines,[8]

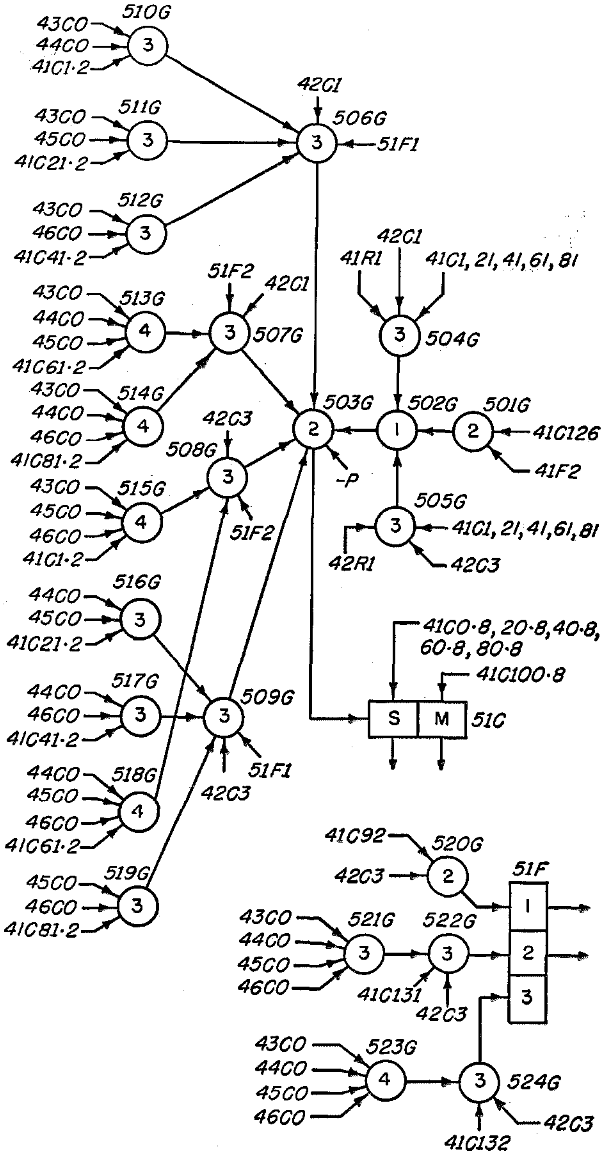

Figure 18—Error-correcting-system character emission, incoming terminal.

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[6] British Patent Application 34415/53.

[7] British Patent Application 34415/53.

[8] British Patent Application 3433/54.

or cathode-ray tubes, each of which would produce a repetition rate of approximately 1 millisecond. An alternative approach is possible with a magnetic drum having a rotational time of, say, 20 to 50 milliseconds. In each case the contents of the memory are scanned by sequential examination of the elements, and as a consequence their condition can only be read or modified at specific instances of time indicated by a special series of pulses produced for this purpose. A group of elements in a memory corresponds closely to the register indicated as *1R* in Figure 1. Each single element can be used to represent a trigger such as *1F*, and with some auxiliary equipment a group of elements can be caused to act as a decimal or binary counter.

The first cost and life of a memory element compares very favourably with any individual component that needs to be packaged, tested, and wired into a circuit. It will be appreciated, therefore, that in making any economic comparisons it is essential to study the associated equipment needed to use the memory to represent a variety of different functions; the timing, or clock, pulses already mentioned form one example. Circuits for writing, reading, and amplifying form another group, and in addition there may be control circuits for performing simple arithmetic. In any case, it will be essential to drive the scanning arrangement. Some of these expenses may be common to a number of groups of elements, whereas others may be individual to a single group like the track on a magnetic drum. The extent of the charges for the common equipment per storage element will naturally depend on the facilities provided and on the number of elements forming the memory. For many applications the number of storage elements will be large and as a consequence the speed of examination will be high, requiring that hard valves or transistors be employed. There is no question that there is likely to be an optimum scanning speed because the expense of the memory as a whole must necessarily be influenced by the power consumption, which will tend to increase rapidly with the frequency of operation.

It is a matter of some interest that a register shown on a functional diagram may in general be a series of static devices, like valves, or a memory. If a memory is arranged to perform a number of different functions as that of a counter, trigger, or register, there arises a need to express the state of the memory by a waveform or as a binary number. This is analogous to the technique of describing pulse supplies by their waveform, but it should be appreciated that, whereas static devices are invariable, memories vary according to the circuit condition.

Cubic Distortions in Ring Modulators*

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THE IDEAL ring modulator shown in Figure 1A with its equivalent circuit at 1B may be thought of as being a polarity-changing switch. If the terminals to which the carrier wave E_c is applied were connected to a battery and the left-hand terminal were positive, the two rectifiers shown on the horizontal lines would be highly conductive and the rectifiers on the diagonal lines would be relatively nonconductive. A signal wave E_s applied at the input would pass through the input transformer, the horizontal rectifiers, and the output transformer to the load. If the battery polarity were reversed, the diagonal rectifiers would be conductive and the horizontal rectifiers would be nonconductive. The signal wave would then pass to the load through the diagonal rectifiers but the polarity of the wave at the load would be the reverse of the previous case.

Assuming the carrier frequency to be several times that of the signal and its waveform and amplitude to be such as to switch the rectifiers rapidly between conducting and nonconducting conditions, a waveform of the type shown in Figure 1C would result.

The frequency spectrum across the load resistor R_l is the product of the signal voltage E_s and the switching function α. Only sidebands

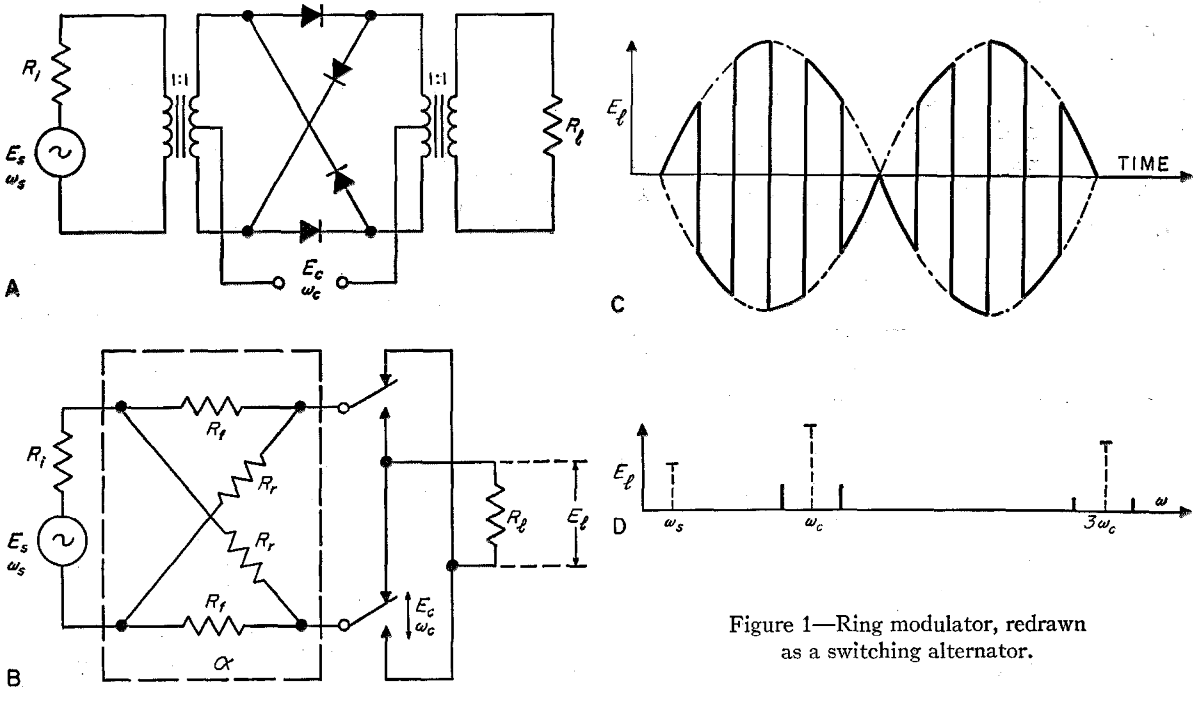

Figure 1—Ring modulator, redrawn as a switching alternator.

$$\alpha = \ln \frac{(R_f/R_r)^{1/2} + 1}{1 - (R_f/R_r)^{1/2}} = \text{attenuation of switching alternator,}$$

$$E_l = \frac{E_s e^{-\alpha}}{2} \cos \omega_s t \times \underline{+1}\big|\overline{-1} = \frac{E_s e^{-\alpha}}{\pi} \cos\,(\omega_c \pm \omega_s)\,t - \frac{E_s e^{-\alpha}}{3\pi} \cos\,(3\omega_c \pm \omega_s)\,t \pm \cdots$$

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* Reprinted from *Frequenz,* volume 5, number 11/12, pages 298–303; 1951. Dedicated to Professor Barkhausen on his 70th birthday.

of the first order of the carrier wave and its odd harmonics will appear as indicated by $3\omega_c$. The deviation of the actual from the ideal case is determined by three factors.

A. Nonideal switching action of the rectifier.

B. Incomplete symmetry of the network.

C. Interaction of signal voltage and rectifier circuit.

In this paper, only the action of the signal voltage on the performance of a modulator in its ideal conception will be analyzed. This action of the signal (even for complete symmetry of the circuit) produces cubic distortion in the modulator. This happens even if the carrier wave has the ideal shape. The most disadvantageous distortion comes from sidebands of the third order of the signal frequency that is, sidebands located at three times the signal frequency from the carrier frequency. These sidebands are particularly unwanted in multichannel telephony because they generate both intelligible and unintelligible crosstalk.

The following considerations are actually limited to ring modulators consisting of rectifier elements with small capacitances (for instance, germanium modulators). It is more complicated to consider rectifiers in which the capacitance varies with the impressed voltages as is the case with copper-oxide and selenium rectifiers.

1. Generation of Cubic Distortion

Since the individual nonlinear elements of the network are subjected to different operating conditions in a ring modulator, it seems appropriate to differentiate between forward and reverse distortions. Forward distortion occurs when a rectifier element is passing current, while reverse distortion originates in elements that block the current.

1.1 Forward Distortion

Figure 2 shows the dependence of the forward (current-passing) resistance R_f of a rectifier element in the signal-voltage circuit on the voltage across the element. The curve shown is the first derivative of the reverse function of the direct-current characteristics $i = f(e)$ of the rectifier element.

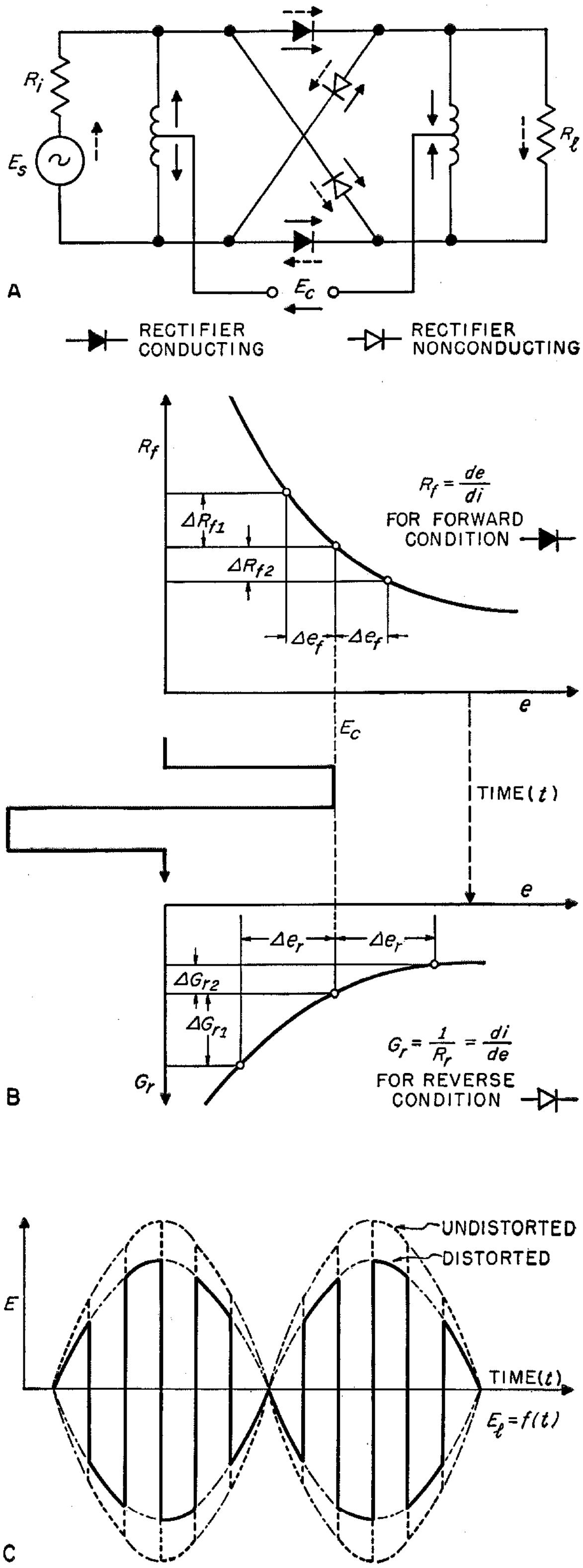


Figure 2—Generation of cubic distortion in conducting and blocking rectifier elements.

The following observations refer to the duration of one switching position, that is, one half-wave of the carrier voltage, which is assumed to be rectangular to simplify the conditions of operation.

The operating point on the line $R_f = g(e)$ is determined by the carrier voltage applied to the element. At insignificantly small signal voltages, the resistance of the element for the signal circuit equals R_f. When the signal voltage rises, it adds to the carrier voltage in one element and reduces the carrier voltage by the signal-voltage value in the other one. The voltage at the element is thus increased and decreased cyclically by Δe_f, in which Δe_f is approximately equal to the product of the signal current and the differentiated resistance of the rectifier element.

Corresponding to the nonlinear relation between R_f and e, the sum of the two rectifier-element resistances forming part of the signal circuit will, hence, change with the instantaneous value of the signal voltage across the element. The shape of the output voltage indicates that the distortion is cubic because the odd harmonics of the envelope, with the polarity changed, point to odd sidebands.

From the graph of Figure 2B, it may be deduced directly that the forward cubic-distortion factor D_{3f} is proportional to the difference in the change of resistance $\Delta R_f = \Delta R_{f1} - \Delta R_{f2}$ of the rectifier element divided by the sum of the resistances ΣR in the signal circuit.

$$D_{3f} \sim \Delta R_f / \Sigma R. \tag{1}$$

Applying Taylor's theorem for the slope of the characteristic $R_f = g(e)$ in the vicinity of the operating point E_c as determined by the carrier voltage,

$$\begin{aligned} R_f - \Delta R_{f2} &= g(E_c + \Delta e_f) \\ &= R_f + \frac{(dR_f)}{(de)}\frac{\Delta e_f}{1!} + \frac{(d^2R_f)}{(de^2)}\frac{\Delta e_f^2}{2!} + \cdots, \end{aligned} \tag{2}$$

$$\begin{aligned} R_f + \Delta R_{f1} &= g(E_c - \Delta e_f) \\ &= R_f - \frac{(dR_f)}{(de)}\frac{\Delta e_f}{1!} + \frac{(d^2R_f)}{(de^2)}\frac{\Delta e_f^2}{2!} - \cdots, \end{aligned} \tag{3}$$

and, by addition of (2) and (3),

$$\begin{aligned} \Delta R_f &= \Delta R_{f1} - \Delta R_{f2} \\ &= \frac{(d^2R_f)}{(de^2)} \Delta e_f^2. \end{aligned} \tag{4}$$

The effects of the higher derivatives of the characteristic are neglected. The forward cubic distortion contributed by the conducting rectifier elements is approximately proportional to the square of the signal-voltage component Δe_f on the conducting rectifier element and directly proportional to the curvature of the $R_f = g(e)$ characteristic at the operating point determined by the carrier voltage applied to the conducting rectifier element. By curvature, the second derivative of the resistance characteristic is meant.

1.2 Reverse Distortion

Analogous considerations apply to the reverse distortion, except that the conductance takes the place of the resistance. Starting with the characteristic of the differential conductance $G_r = (1/R_r) = f(e)$, it will be found that the distortion factor is proportional to the change in conductance in relation to the total conductance in the signal circuit. The results are analogous to (4).

$$\Delta G_r = \frac{(d^2G_r)}{(de^2)} \Delta e_r^2. \tag{5}$$

The cubic-distortion factor contributed by the blocking rectifier elements also varies as the square of the signal-voltage component on the blocking rectifier element Δe_r and the curvature of the characteristic $G_r = f(e)$ at the operating point, determined by the carrier voltage applied to the blocking rectifier element.

As will be shown later, the signal-voltage component on the blocking rectifier element is, however, very different from that on the conducting rectifier element.

It may be seen readily that the distortions in both the conducting and the blocking elements have the same direction when the rectifier element in its blocking period acts in accordance with the characteristic of differential conductance indicated in Figure 2B. This is the case with germanium diodes. Barrier-layer rectifiers have different curvatures for the blocking period. In germanium modulators, therefore, the distortion factors will have to be added.

All endeavor to decrease the distortion in ring modulators must, hence, aim at the reduction of two quantities; the signal voltage applied to the rectifier element and the degree of curvature of the characteristic for resistance and conductance, respectively, which is influenced by the carrier voltage.

2. Signal-Voltage Component

The signal-voltage component that appears across the rectifier element during one period of a switching alternation results from the additional current sent through the rectifier-element resistance by the signal-voltage source. It is, therefore, nearly equal to the product of the signal current and the differential resistance at the operating point set by the carrier. The signal-voltage component may then be computed from the signal-voltage distribution across the line and cross resistances of the modulator.

Assuming a large switching ratio of the rectifier element, that is $R_f < R_l < R_r$, the following equations are obtained.

$$\Delta e_f \approx \frac{R_f}{R_i + R_l} E_s, \tag{6}$$

$$\Delta e_r \approx \frac{R_l}{R_i + R_l} E_s, \tag{7}$$

where

R_i = internal resistance
R_l = load resistance
E_s = terminal voltage of the signal circuit.

The quantities Δe_f and Δe_r are proportional to the terminal voltage of the signal circuit, provided the carrier voltage and the circuit of the network are constant. Indeed, all measurements on ring modulators always show the square-law character of the cubic-distortion factor predicted by (4) and (5).

From (6) and (7), it follows that the signal voltage across the blocking rectifier element of a ring modulator is higher than across the conducting element by the factor

$$R_l/R_f = (R_r/R_f)^{1/2},$$

in other words by the square root of the switching ratio. This equation explains the fact, established by measurements, that in the case of germanium modulators, too, the reverse distortion is always higher than the forward distortion in the matching condition. It is because of this fact that a controlled mismatching of the ring modulator, called undermatching, is effective as a means for linearizing the action.

3. Curvature of Resistance Characteristic and Conductance Characteristic

The curvature of the related characteristics for resistance and conductance depends on the type of rectifier. If the direct-current characteristic $i = f(e)$ can be represented by a parabola, then the curvature of the resistance characteristic $R_f = g(e)$ follows the equation $T \sim 1/e^3$, where T is the degree of curvature. The cubic-distortion factor thus varies inversely with the cube of the carrier voltage. With a direct-current characteristic in the form of an exponential function, the curvature of the resistance characteristic and, consequently, the cubic-distortion factor will decrease as an exponential function together with carrier voltage. In the first case, there is a linear relation between the distortion attenuation and the carrier level and in the second case between the distortion attenuation and the carrier voltage. Here, distortion attenuation stands for the natural logarithm (ln) of the reciprocal distortion factor.

Measurements on germanium modulators have shown that distortion attenuation increases in proportion to nearly three times the carrier level when the carrier voltage is low, and nearly to twice the carrier level when the carrier voltage is higher. It is therefore evident that the differential-resistance characteristic $R_f = g(e)$ must be a hyperbola for low carrier voltages and a negative logarithm of the form $R_f = C - \ln_e$ for higher carrier voltages. From this information on the resistance characteristic, it follows further that the direct-current characteristic of the germanium diode can be plotted as a parabola only when the voltage is small and that its curvature is less pronounced when the voltage is increased. This is confirmed by measurements on direct-current characteristics.

4. Linearization of a Modulator

The effectiveness of all methods employed to reduce cubic distortion in ring modulators can easily be predicted by the above considerations. It is only necessary to determine the effect the applied method will have on the signal-voltage component across the rectifier element in the conducting and blocking conditions and on the setting of the operating point, determined by the carrier voltage, for both the conducting and blocking conditions of the rectifier element. Then, the effects on the forward and reverse distortions may be analyzed separately and the over-all results can be qualitatively estimated. It may be stated that a nonlinear network of the ring modulator is linearized when the distortion attenuation has increased, while the output power level and the carrier power level have remained constant.

Data on distortion attenuation that are supported by voltage-level values only are not sufficient to define a degree of distortion suppression. Linearizing action frequently results in the distortion remaining nearly unchanged at constant carrier voltage, while the carrier input resistance is increased. An improvement can only be obtained when the carrier voltage is increased so as to yield the original carrier power. An unequivocal criterion for the nonlinear action of the modulator is the graph of distortion attenuation plotted against the carrier power level for a predetermined power level at the modulator output terminals.

5. *Measurement Results*

Some of the best-known methods for linearizing action were tested experimentally and will be discussed against the background of the above nonlinear relations. All measurements were made on germanium rectifiers. It should be noted that the condition of mathematically precise square-wave carrier voltage could not be fulfilled and that the terminal resistance, too, was not a real one for all frequencies but consisted of the input resistance of a filter impedance. This departure from theoretical conditions made it impossible to obtain exact verification of the factors causing distortion. Nevertheless, the quality of various relations can be clearly observed, as shown by the results. For the purpose of the discussion, however, a graph of the distortion attenuation is plotted against both the carrier voltage and the carrier power; the first provides for the application of the above-described considerations under conditions of constant carrier voltage, while the second points to an effective method of linearization. Reference level of zero nepers is 1 milliwatt in 600 ohms (0.7749 volt).

Figure 3 shows the effect of inserting a resistor in series with each rectifier element. The distortion in the blocking condition is unchanged for a constant carrier voltage applied to the modulator. The distortion in the forward direction is increased despite the fact that the series resistance R_s reduces the signal voltage across the rectifier element, because it also reduces the carrier voltage and thus sets the operating point on a more-curved portion of the characteristic. The two effects do not entirely cancel each other. This is especially true for large values of series resistance, where the distortion due to the reduction of the carrier voltage across the element predominates and the curvature

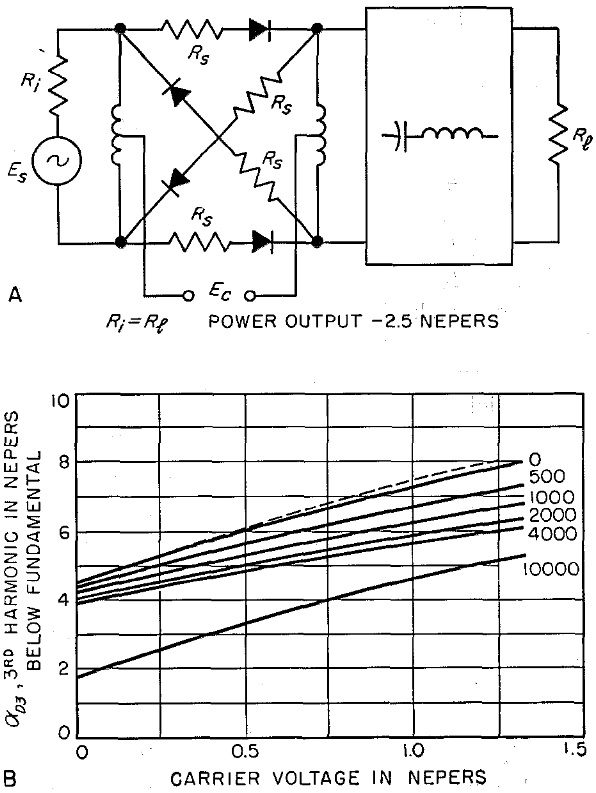


Figure 3—Measured distortion attenuation in a germanium modulator as a function of the carrier voltage level for the indicated values of R_s in ohms. The dashed curve is for 10 000 ohms in parallel with the rectifiers.

increases, as discussed in the previous section. This increased curvature, the carrier voltage remaining low, is clearly seen from the slope of distortion attenuation plotted against the carrier voltage level, when R_s equals 0. If a low carrier voltage is applied to the modulator, the series resistance has little effect since the voltage drop from the carrier current is correspondingly small. The distortion, however, has decreased as may be seen when the curve of attenuation is plotted against the carrier power as in Figure 4. This is accounted for by the increase in carrier input resistance due to the added series resistor.

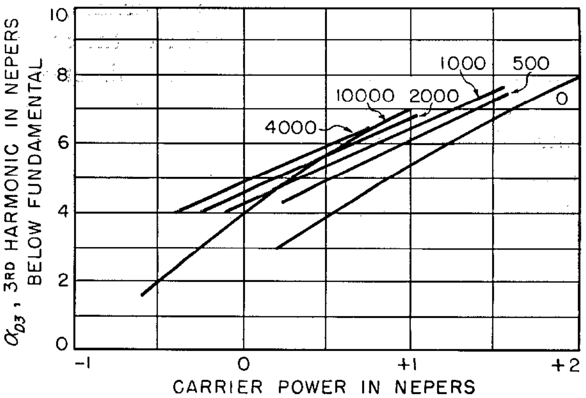


Figure 4—Measured distortion attenuation of the circuit of Figure 3 as a function of carrier power level for the indicated values of R_s in ohms.

The network of Figure 3 is not always suitable because the over-all attenuation of the modulator increases rapidly. A resistance R_p placed in parallel with the rectifier elements does not change the distortion relations so long as this resistance is high with respect to the load resistance, because no change then occurs in the factors that determine the distortion. Similarly, there is no change in the carrier power requirements.

An arrangement commonly used for linearization is shown in Figure 5. The series resistors R_s are inserted at the signal-input side; they are outside the rectifier network. This circuit differs from the previous one in that the voltage across the blocking rectifier element is reduced by the drop due to the current flowing through the resistance of the conducting rectifier. This is also true for the signal voltage at the blocking element. However, the two changes again do not completely cancel each other. If the carrier voltage is kept constant, the distortion attenuation decreases. The result of this linearizing action would again be most evident from a graph of the attenuation plotted against the carrier power and would not differ markedly from Figure 4. Similarly, this circuit is not too practical because of the substantial increase in attenuation of the entire modulator output.

Figure 6 is for a split-ring arrangement. In this modulator, the four rectifier elements do not constitute a closed ring but two push–pull systems supplied from separate windings of the input transformer. Both half-modulators get the carrier voltage through separate compensating resistors R_s, the values of which were varied for the measurements.

If the carrier voltage is maintained constant, the distortion due to the reverse currents through the blocking rectifier elements remains unchanged even though the resistor values are varied. The distortion attenuation caused by the conducting

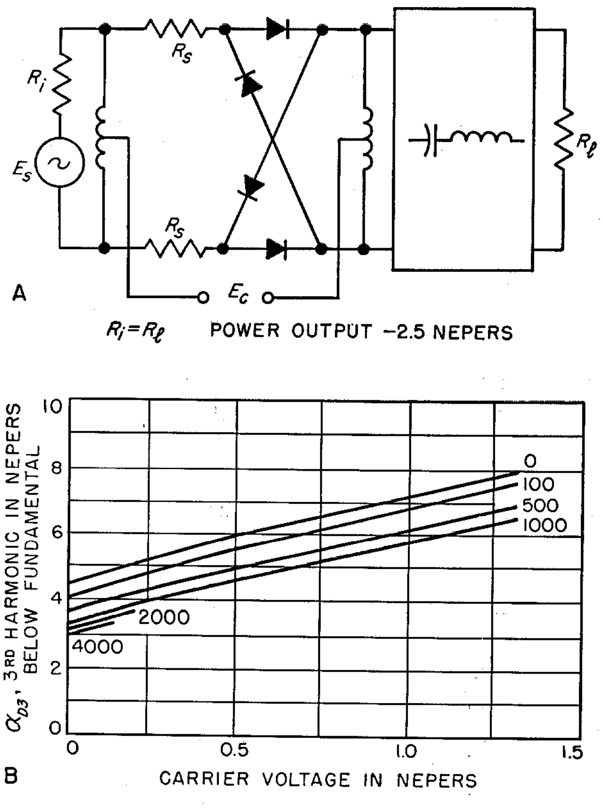

R_i
R_S
R_S
E_S
R_ℓ
E_C
A
$R_i = R_\ell$ POWER OUTPUT -2.5 NEPERS
α_{D3}, 3RD HARMONIC IN NEPERS BELOW FUNDAMENTAL
10
8
6
4
2
0
0
0.5
1.0
1.5
B
CARRIER VOLTAGE IN NEPERS
0
100
500
1000
2000
4000

Figure 5—Measured distortion attenuation as a function of carrier voltage level for indicated values of R_s in ohms.

rectifier elements, however, decreases noticeably as the value of R_s is increased. This should be expected because the reduction in the carrier voltage at the conducting rectifier element is not compensated for by a corresponding reduction in signal voltage at the element. Here, too, the attenuation plotted against carrier power (Figure 7) shows a marked effect of the linearizing action.

This example shows the principle on which all discussed systems of linearization are based: the reverse distortions are prevalent in germanium rectifiers for matched loads. The conducting rectifier element contributes only a small proportion of the total distortion. In view of the above discussion, it is possible to increase the forward distortion somewhat by the insertion of input resistances that reduce the carrier voltage across the conducting element. Since only the conducting element determines the carrier power requirements, it is possible to save carrier power, which may be used to increase the total carrier voltage at the modulator. The improvement in linearization brought about by increasing the carrier voltage more than compensates for the increase in forward distortion. The linearizing action, then, consists in a more-appropriate distribution of all distortion among conducting and blocking elements. If this process is exaggerated, the forward distortion will increase more rapidly than the reverse distortion decreases, and the total distortion will increase. Therefore, the attenuation plotted against carrier power for various values of the parameter R_s shows an optimum value of R_s.

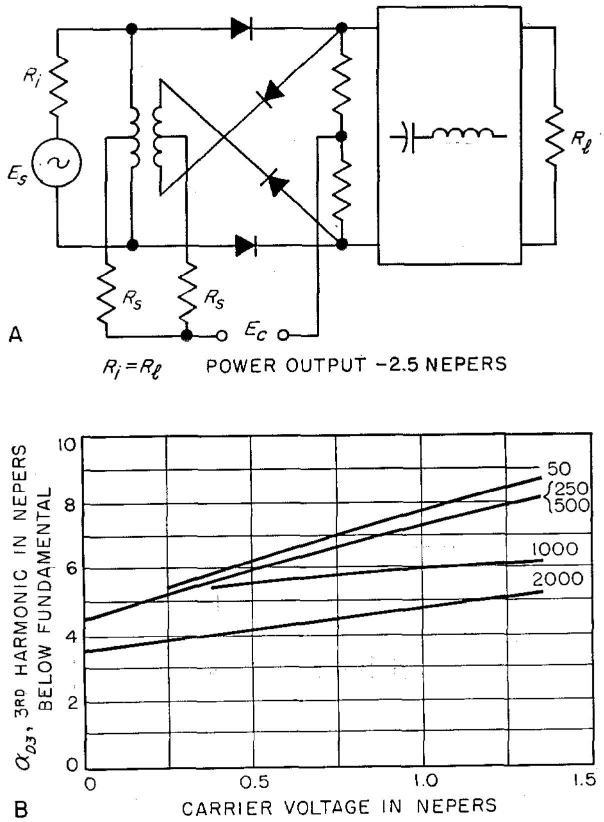


Figure 6—Measured distortion attenuation as a function of carrier voltage level for various values of R_s in ohms.

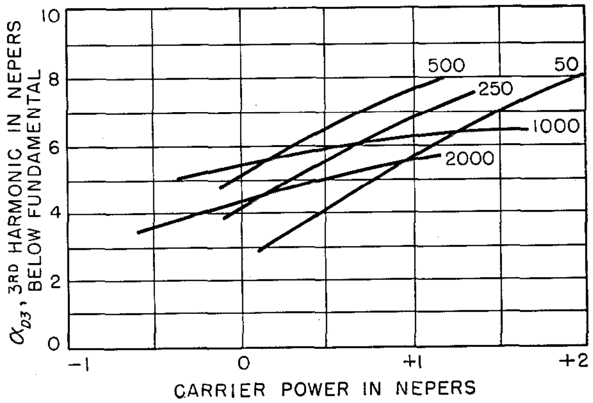


Figure 7—Measured distortion attenuation of the circuit of Figure 6 as a function of carrier power level for values of R_s in ohms as indicated.

In the split-ring circuit, the advantage stems from the fact that the carrier voltage applied to the conducting element is not reduced at the expense of modulator attenuation because the resistance R_s is not in the signal circuit. The attenuation is raised by only the small amount attributable to the change in switching ratio.

Another method of distortion suppression in ring modulators is based on the value of the load resistance R_l. As is evident from (6) and (7), the forward and reverse distortions are affected oppositely by changes in R_l but not to the same degree. Since the distortions vary as the square of the signal voltage applied to the rectifier element, the sum of these two distortions will also vary with the load resistance. Maximum distortion attenuation will be obtained when both types of distortion are equal. Results of measurements on germanium modulators show that this maximum occurs with undermatching, from which it is to be concluded that at the proper matching condition the reverse distortion predominates. Proper matching is defined as the condition in which the internal resistance R_i of the signal-current source, the load resistance R_l, and the input and output resistances of the modulator ring are all equal. Undermatching is the condition in which the load resistance is of a smaller value than the other resistances.

Figures 8 and 9 show the measurement results for a split-ring modulator. The circuit in Figure 8 has a resistor as R_l, while R_l in Figure 9 is a filter circuit. For the resistive load of Figure 8, the qualitative laws can still be traced; this is not so for Figure 9, where the signal-voltage component across each element can be predicted only when an impedance curve of the filter circuit is available. It is clear, however, that in the more-common practice of a reactive load, a better linearizing action is possible than with a resistive load.

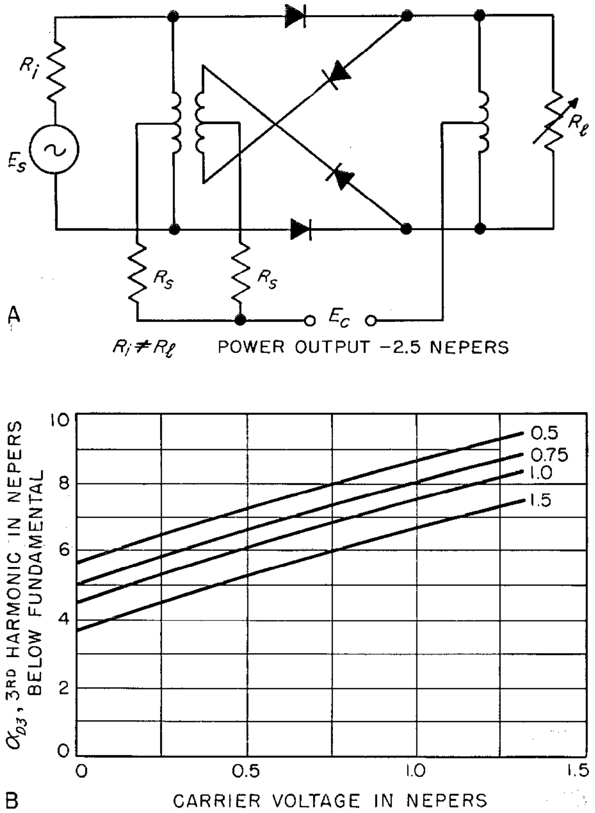
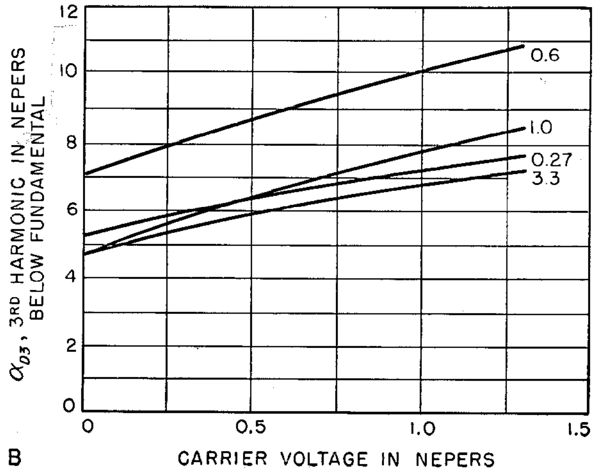
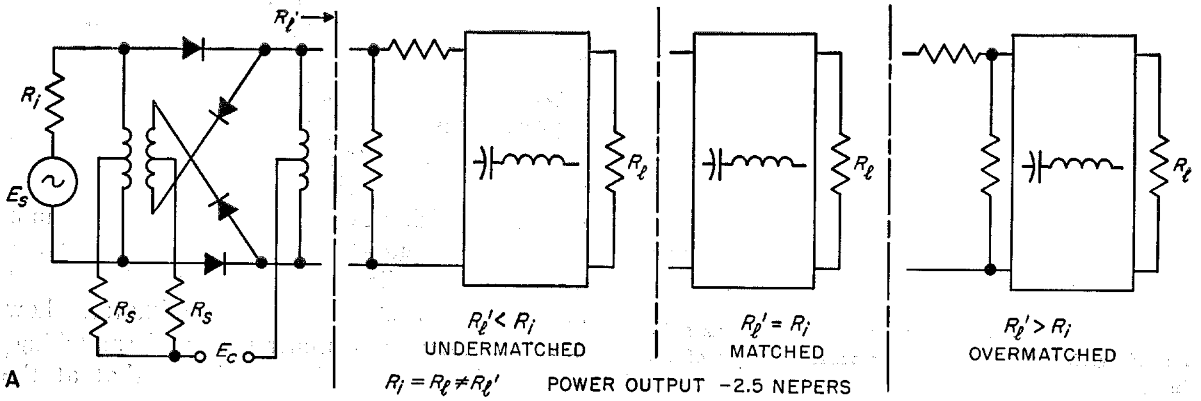


Figure 8—Above, measured distortion attenuation as a function of carrier voltage level for the values of R_l/R_i indicated. R_s = 400 ohms.

Figure 9—At right and below, measured distortion attenuation as a function of carrier voltage level for the indicated values of R_l'/R_i. R_s = 400 ohms.





The optimum load resistance will, naturally, depend on all parameters that can influence the forward and reverse distortions; both should be of equal value for best results.

In practice, it will often be necessary to use several of the methods discussed here simultaneously because the circuit is affected by other factors, too. For instance, the degree of efficiency or tolerances for a required symmetry of the transformers may influence the design. The circuit of Figure 3, for instance, might be improved by providing a more-suitable terminal resistance; or, the arrangement of Figure 6 may be expanded by inserting resistances in series with the rectifier elements. The results of measurements show, moreover, that any distortion in a ring modulator that cannot be affected by circuit symmetry, can be linearized by up to 5 nepers through appropriate circuit considerations. This value refers to the distortion attenuation of the matched, not-linearized, modulator at constant signal and carrier power levels. For zero power levels at the modulator output terminals and the carrier input terminals, the value of the attenuation a_{D3} is about 2.5 nepers.

Data for suppression of distortions at other signal levels are derived from the square-law equation of signal level and cubic distortion. The distortion attenuation increases at about twice the rate of signal-level reduction. For germanium rectifier elements operated at approximately zero level, the distortion attenuation increases at about three times the rate of carrier-level reduction. For ordinary carrier-telephony service with the customary carrier power level of +0.7 neper and a usually required output level of −3.5 nepers, it is possible to obtain distortion attenuation of more than 11 nepers.

Since the methods of linearizing action discussed here are designed to affect the generation of distortion in the nonlinear rectifier element, they cover not only cubic distortions but also tend to suppress the squared modulation output, especially the second-order sidebands that are already reduced by the symmetry of the ring-modulator circuit.

Microstrip Applied to Band-Pass Microwave Filters*

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MICROSTRIP[1–3] is a wide-band transmission system in which electromagnetic waves are propagated through a dielectric medium bonded by a strip conductor on one side and a conducting ground plane on the other side as is illustrated in Figure 1.

A transverse electromagnetic mode can propagate in microstrip only if the surrounding dielectric is uniform and infinite and the system is lossless. However, practical microstrip lines may consist of a solid dielectric on which a conductive strip is printed; there is an air region above them. Since these lines involve composite dielectrics, they cannot support a pure transverse electromagnetic mode. Nevertheless, both theory and experiments indicate that the field and power flow are concentrated in the dielectric between the conductors and that the assumption of a single infinite dielectric leads to useful results even though it is not rigorous. A first-order theory based on this assumption was described in a previous paper.[2]

The present paper deals with an experimental investigation of the principal characteristics of microstrip. The phase velocity and attenuation of the dominant mode have been measured for various geometries of the lines over a wide frequency range. The characteristics of various types of obstacles in the line, as defined by their scattering-matrix coefficients or equivalent circuits, have been determined and are shown in some examples. From the values of the susceptances of obstacles and the characteristics of the lines, a resonant section can be designed. This design theory agrees well with the results of an experimental investigation.

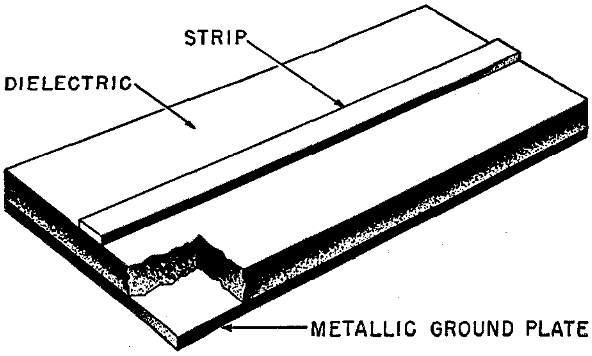

Figure 1—Microstrip line.

1. Characteristics of Microstrip

1.1 EXPERIMENTAL MEASUREMENTS

Since the geometry of microstrip is rather unconventional, the usual measurement techniques must be modified to obtain the phase velocity and attenuation characteristics of the line or the characteristics of obstacles in the line. For example, a probe can be used only to obtain information on the field distribution in a region far from the main power flow. Also a slotted section probing through the ground plane under the strip conductor can alter the normal mode of propagation of energy.

For these reasons, it has been found convenient to make all impedance measurements by connecting the microstrip line through a transducer to a waveguide or to a coaxial slotted line.

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* Abridgment of a paper, "Characteristics of Microwave Printed Lines (Microstrip) with Application to the Design of Band-Pass Microwave Filters," presented at the joint meeting of the Washington Section of the Institute of Radio Engineers and the American Section of the International Scientific Radio Union at Washington, District of Columbia, on April 23, 1952.

[1] D. D. Grieg and H. F. Engelmann, "Microstrip—A New Transmission Technique for the Kilomegacycle Range," *Electrical Communication*, volume 30 pages 26–35; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1644–1650; December, 1952.

[2] F. Assadourian and E. Rimai, "Simplified Theory of Microstrip Transmission Systems," *Electrical Communication*, volume 30, pages 36–45; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1651–1657; December, 1952.

[3] J. A. Kostriza, "Microstrip Components," *Electrical Communication*, volume 30, pages 46–54; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1658–1663; December, 1952.

The reflections from the microstrip line are then measured in the standard manner. After correction for the effect of the junction, the parameters necessary to determine the various characteristics of the microstrip line can be obtained. Although the method has been described originally by Deschamps[4–8] for the determination of the reflection coefficients and attenuation losses of a closed waveguide junction, its application to microstrip is justified as long as the main power flow along the line is confined in one dominant mode, and in general this has been found to be true. A brief resume of Deschamps' method is given in the appendix.

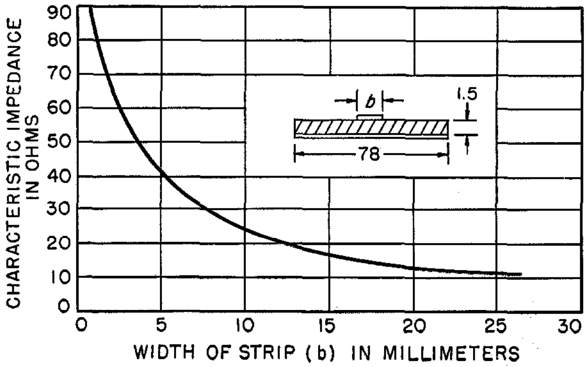

Figure 2—Characteristic impedance for microstrip as a function of the width of the strip. Fiberglas *G-6* dielectric is used. Dimensions are in millimeters.

1.2 Propagation Constants

The following results have been obtained experimentally with microstrip.

A. Over a wide range of frequencies, the phase velocity v is constant to within experimental errors that are less than 0.5 percent; that is, microstrip is nondispersive.

B. In general, the wavelength in microstrip is higher than the value λ_0 corresponding to a transverse electromagnetic mode of propagation between infinite parallel plates. For a given thickness of dielectric, the wavelength decreases as the width of the strip conductor increases, approaching asymptotically the value λ_0 obtained for very wide strips.

C. The *dominant mode propagating in microstrip* is not a pure transverse electromagnetic mode. Consequently, the definition of the characteristic impedance of the line does not have the same meaning as it would have in a coaxial system. However, to a first approximation, a characteristic impedance could be defined as

$$Z_0 = 1/Cv, \qquad (1)$$

where C is the measured electrostatic capacitance per unit length of microstrip and v the phase velocity along the microstrip. The characteristic impedance as a function of the width of the strip is shown in Figure 2. For purposes of comparison with the theoretical curves presented[2] earlier, the experimental results of the ratio of Z_0/Z_0' versus b/h have been plotted in Figure 3.

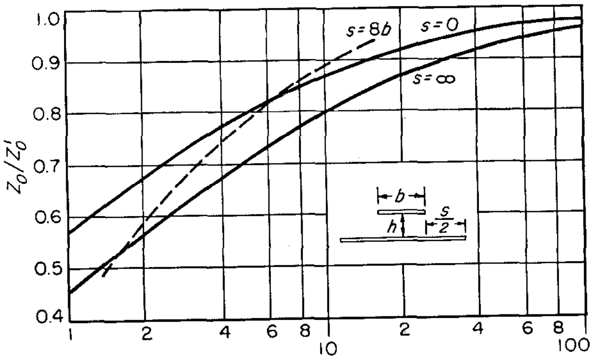

Figure 3—Theoretical limits of the characteristic impedance for a wide strip of zero thickness above a finite ground plane are shown by solid lines. Experimental values are given by the broken-line curve.

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[4] G. A. Deschamps, "Application of Non-Euclidean Geometry to the Analysis of Waveguide Junctions," presented at the joint meeting of American Section of the International Scientific Radio Union and the Institute of Radio Engineers in Washington, District of Columbia, on April 23, 1952. Published in part in footnote references 5–8.

[5] G. A. Deschamps, "Determination of the Reflection Coefficients and Insertion Loss of a Waveguide Junction," *Electrical Communication,* volume 31, pages 57–62; March, 1954: also, *Journal of Applied Physics,* volume 24, pages 1046–1050; August, 1953.

[6] G. A. Deschamps, "Geometric Viewpoints in the Representation of Waveguides and Waveguide Junctions," *Proceedings of the Symposium on Modern Network Synthesis,* Polytechnic Institute of Brooklyn, New York; September 30, 1952: pages 277–295.

[7] G. A. Deschamps, "New Chart for the Solution of Transmission-Line and Polarization Problems," *Electrical Communication,* volume 30, pages 247–254; September, 1953: also, *Transactions of the IRE Professional Group on Microwave Theory and Techniques,* volume 1, pages 5–13; March, 1953.

[8] G. A. Deschamps, "Accurate Comparison of High Standing-Wave Ratios and Its Application to the Determination of the Attenuation Constant of a Waveguide," presented at the 3rd Conference on Ultra-High-Frequency Measurements in Washington, District of Columbia, on January 16, 1953: unpublished.

Z_0' is the characteristic impedance corresponding to a constant field in the absence of fringing and leakage flux $Z_0' = (h/b)(\mu/\epsilon)^{1/2}$. It can be seen that the agreement with the first-order theory is quite good.

D. It was found that the attenuation along the microstrip line is constant with the exception of a small region near the transducer. Also, for practical lines, the experimental value of the attenuation constant of the dominant mode agrees closely with the value given[2] by the first-order theory. In the case of 50-ohm microstrip lines printed on Teflon-impregnated Fiberglas $\frac{1}{16}$ inch (1.6 millimeters) thick, the Q of the line is about 560 at 6000 megacycles per second, corresponding to an attenuation of 1.4 decibels per meter.

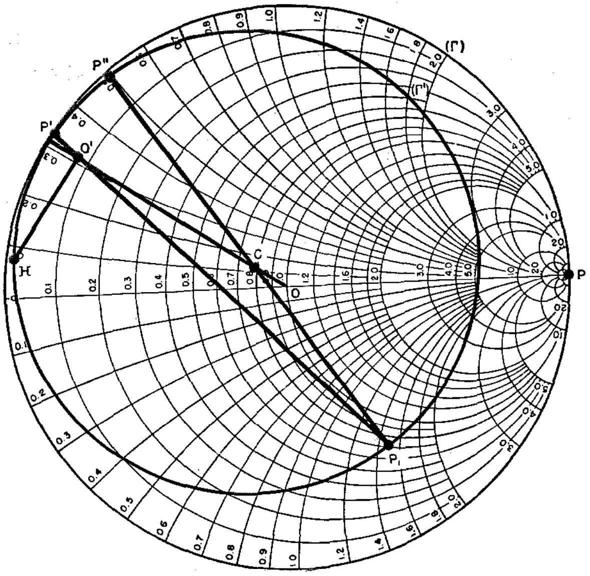

Figure 4—Circle diagrams for a transverse post in microstrip. $S_{11} = -0.725 + j0.49$, $\pm S_{12} = 0.232 + j0.437$, and $S_{22} = -0.790 + j0.303$.

2. *Characteristics of Obstacles*

The characteristics of various obstacles in microstrip have been studied,[9] but only two examples will be described here.

2.1 STEP DISCONTINUITY

Experiments show[9] that the equivalent circuit of a microstrip step discontinuity is an ideal transformer if the losses in the line are small. In a first approximation, the square of the turns ratio n^2 of the equivalent transformer is equal to Z_2/Z_1, where Z_2 and Z_1 are the values of the characteristic impedance of the lines on either side of the discontinuity.

2.2 TRANSVERSE POSTS

At a given frequency, the susceptance resulting from a transverse post in the line is a function of the diameter of the post and the location of it across the line.[9] Due to radiation losses, which can be appreciable for large post diameters, it is desirable to specify the susceptances by the scattering-matrix coefficients in which the reference plane is taken as the axis of the post. As shown[4–8] by Deschamps, the power and phase relations expressed by the scattering coefficients can be deduced from the circle diagram shown in Figure 4. In this diagram, circle Γ' of radius R is the image of the unit circle, O' is the image of O, and P' is the image of P through the post. This circle diagram can be obtained by moving a short-circuit behind the post and by measuring for each position of the short-circuit the reflection coefficient as referred to the axis of the post. A good short-circuit is obtained by soldering a large conductive plate between the strip conductor and the ground plane at the end of the line. Other methods for obtaining the circle diagram without destroying the microstrip sample can be found in the references listed in footnotes 4 through 8 and in the appendix.

Introducing the notation $(x)_{\text{db}}$ to mean $-20 \log_{10} |x|$, the insertion loss of the post is given[4–8] by

$$(S_{12})_{\text{db}} = (O'H)_{\text{db}} - (R)_{\text{db}}/2. \qquad (2)$$

An exact equivalent circuit, including the radiation loss in the post, can be derived[4–8] from the circle diagram. When the losses are small, an equivalent shunt inductive susceptance can be deduced (see Figure 5). For large values of susceptances, a frequency-resonant method may be easier to use to determine the values of the scattering-matrix coefficients, as will be shown later.

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[9] M. Arditi, "Experimental Determination of the Properties of Microstrip Components," *Electrical Communication*, volume 30, pages 283–293; December, 1953: also, *Convention Record of the 1953 IRE National Convention*, Part 10, pages 27–37.

Knowing the scattering-matrix coefficients and assuming that the reflection factor measured at the output side of the obstacle is W_2, the transformed reflection factor W_1, which appears on the input side, can be computed from the equation

$$W_1 = S_{11} + \frac{S_{12}^2 W_2}{1 - S_{22} W_2}. \qquad (3)$$

3. Design of Resonant Sections

By placing two obstacles at the proper distance, a microstrip section that is resonant at a given frequency can be obtained.

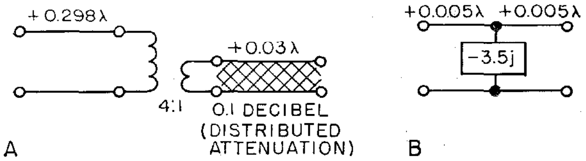

Figure 5—*A* is the exact equivalent circuit including losses deduced from Figure 4 and *B* is the approximate equivalent circuit neglecting the losses for a post transverse to a microstrip line.

The use of wave matrix for the analysis of microwave filters has already been described[10,11] and could be readily applied here. However, as an illustration of Deschamps' methods, the analysis will be carried out here through simple geometrical constructions, thus avoiding to a large extent the use of complicated equations. It will be assumed that the scattering-matrix coefficients of the obstacles are independent of frequency in the pass band of the filter and also, because of the nondispersive properties of microstrip, the angular length of the line between the two obstacles will be made a direct function of frequency.

The analysis will now proceed by steps to show the transitions with previously well-known constructions.

A. Assuming first that the obstacle is a lossless susceptance of normalized value $-jb$, in the Smith chart the image of the unit circle Γ through the obstacle is the unit circle itself, O' is the image of O representing the reflectance of a matched load, and P' is the image of P. (The reference plane is taken at the axis of the obstacle, which is assumed to be infinitely thin; see Figure 6.) Assuming, furthermore, that these points do not change with frequency (a valid assumption over a short frequency range), the locus of the input reflectance at a given point in the line at a distance θ from the obstacle (terminated into a matched load) is a circle Γ_2.

B. By placing two such obstacles at a distance L from each other (see Figure 7), the locus of the input reflectance versus frequency is a circle Γ_3, the image of circle Γ_2 through the second obstacle. This circle can be obtained easily through simple transmission-line theory. At some frequency, the circle Γ_3 passes through the center O of the Smith chart (resonance point). The center C of circle Γ_3 is on the line $O'OB$ and is defined by the following relations shown in Figure 7.

$$OC = OH/2 \qquad (4)$$

with

$$HF/HG = (O'F/O'G)^2. \qquad (5)$$

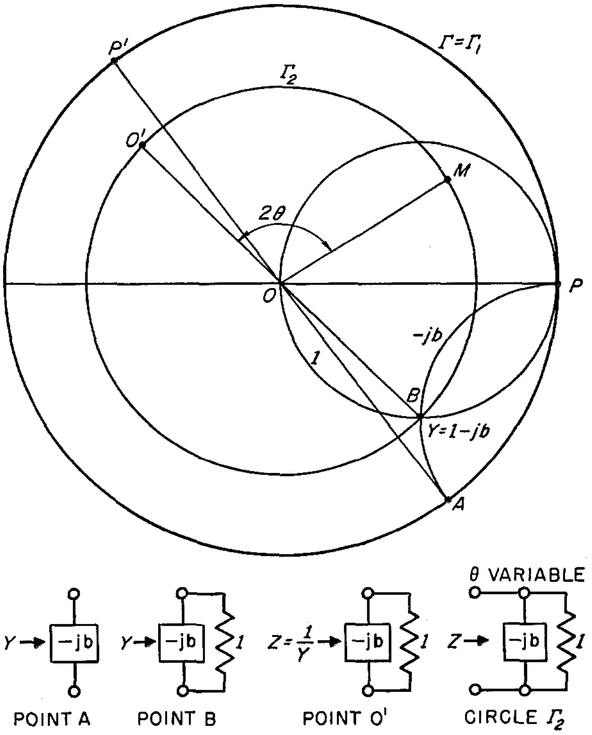

Figure 6—Smith chart representation of the input reflectance of a shunt susceptance across a matched load.

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[10] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Company, New York, New York; 1948: pages 551–554.

[11] J. Hessel, G. Goubau, and L. R. Battersby, "Microwave Filter Theory and Design," *Proceedings of the IRE*, volume 37, pages 990–1000; September, 1949.

At a given frequency, the reflection coefficient is given by the vector *OD* and the transmission coefficient by the vector *ED* orthogonal to *OD*.

For large values of the susceptance b, the circle Γ_3 has a radius almost equal to half the radius of the unit circle (see Figure 8). On this circle, the points marked 1, 2, 3, . . . correspond to the input reflectance at various frequencies. Also in the case of large values of b, the points *E*, *H*, and *O′* almost merge with the point-*G* intersection of the line *OO′* with the unit circle. Moreover, it can be shown[12,13] that the projections from this point of 1, 2, 3, . . . on an axis perpendicular to the line *OO′* give a linear scale of frequencies 1′, 2′, 3′, . . . Thus, the transmission characteristics will be known at any frequency if measurements of the input reflectance are made at only two values of frequency. Also in this case, certain well-known[10] equations are obtained. At resonance,

$$b \tan \theta = -2, \tag{6}$$

3-decibel bandwidth,

$$\Delta f = \frac{4}{\pi}\frac{f_0}{b^2}, \tag{7}$$

and the double-loaded Q_L of the resonant section,

$$Q_L = \pi b^2/4. \tag{8}$$

The input voltage standing-wave ratio and the transmitted and reflected powers can be deduced from this diagram and have been plotted in Figure 9 for the particular case of $|b| = 3$.

It can be seen that when the susceptances b are lossless, the curves are symmetrical. This is not always the case if losses are present.

C. When the resonant section is made with two obstacles that are lossy and in the general case are nonreversible, the frequency response and insertion loss can be obtained through similar geometrical constructions, which have been derived[4–8] by Deschamps and provide some simplification over the analytical computations.

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[12] H. M. Barlow and A. L. Cullen, "Microwave Measurements," Constable and Company, Limited, London; 1950.

[13] W. Altar, "Q Circles, a Means of Analysis of Resonant Microwave Systems," *Proceedings of the IRE*, volume 35, pages 355–361 and pages 478–484; April and May, 1947.

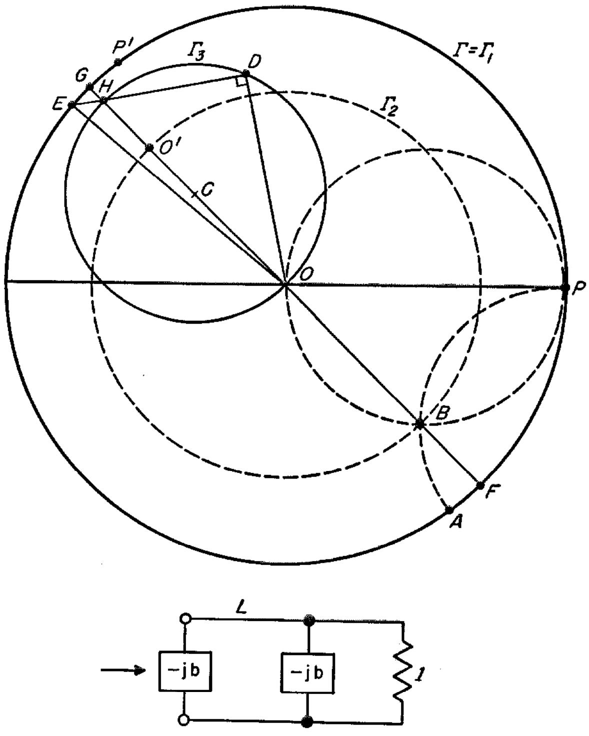

Figure 7—Frequency response of a resonant section made with two lossless susceptances.

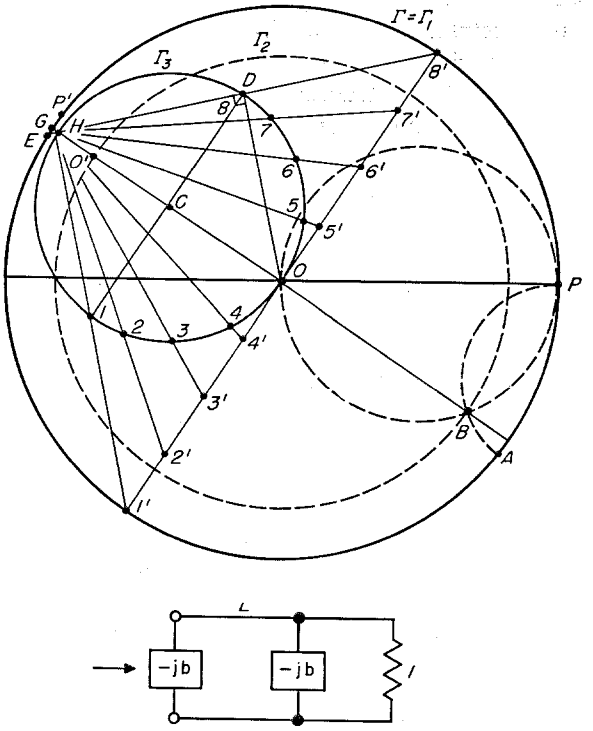

Figure 8—Frequency response and Q of the resonant section for large values of lossless susceptances ($|b| = 3$).

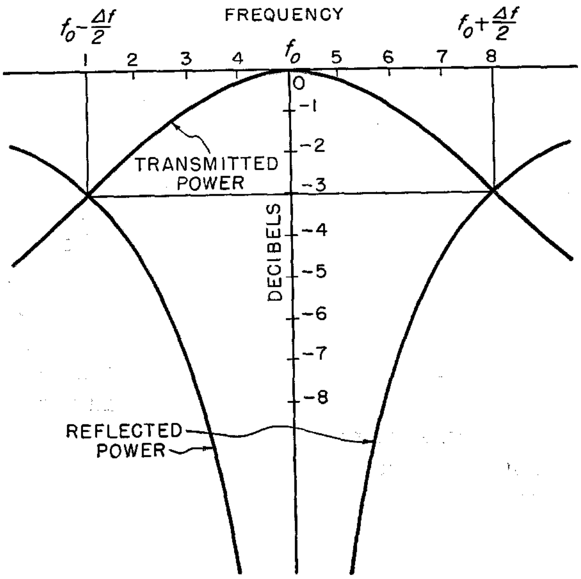

Figure 9—Reflected and transmitted power passing through a resonant section for the case of a lossless susceptance $|b| = 3$.

In Figure 10, the obstacle A is defined by its scattering-matrix coefficients S_{11}, S_{12}, and S_{22} or by the circle Γ_1, the image of Γ through the obstacle; the point O', the iconocenter or image of O; and the point P', the image of P. The locus of the input reflection coefficient when the frequency is varied is a circle Γ_3, the image of circle Γ_2 through the obstacle. Points of the circle Γ_3 can be obtained by using the following relations[5] deduced from (3). The reflection coefficient is

$$\mathbf{OM} = W = S_{11} + \frac{S_{12}^2 S_{11} \exp\left[-2j\theta\right]}{1 - S_{11}S_{22} \exp\left[-2j\theta\right]}, \quad (9)$$

and the transmission coefficient,

$$t = \frac{S_{12}^2 \exp\left[-j\theta\right]}{1 - S_{11}S_{22} \exp\left[-2j\theta\right]}, \quad (10)$$

with

$$\theta = 360^\circ \, Lf/constant.$$

The circle Γ_3 can also be constructed very simply with Deschamps' hyperbolic protractor.[7,14]

At any given frequency,

$$\text{dissipated power} = 1 - |W|^2 - |t|^2. \quad (11)$$

The dissipated power can be computed from the

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[14] G. A. Deschamps, "Hyperbolic Protractor for Microwave Impedance Measurements and Other Purposes," International Telephone and Telegraph Corporation, New York, New York; 1953.

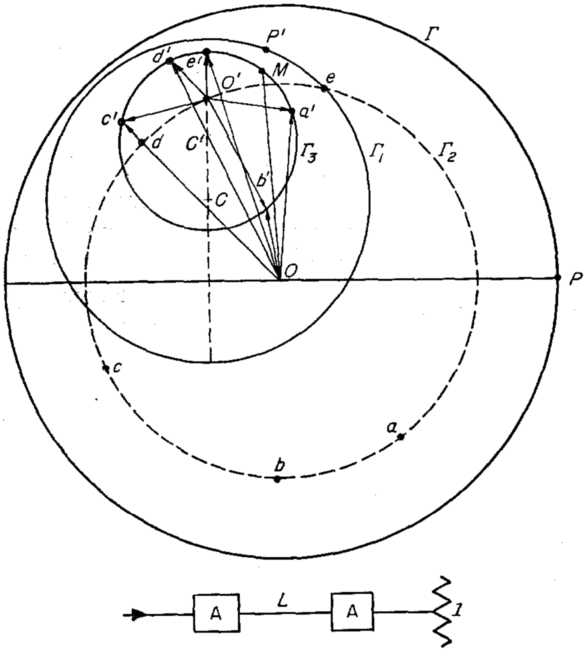

Figure 10—Frequency response of a resonant section with lossy nonreversible obstacles.

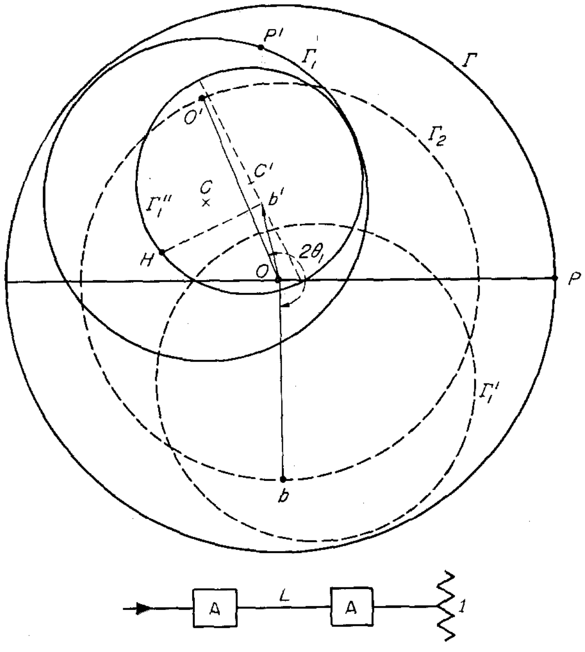

Figure 11—Insertion loss of a resonant section with lossy nonreversible obstacles.

above equations, but here again simple geometrical constructions[5] can give this value easily. The vector **OM** gives W directly. The transmission coefficient t is given by the following construction shown in Figure 11. At the frequency f_1 (point b) for which

$$\theta_1 = 360° \; Lf_1/constant,$$

the circle Γ_1 is rotated around point O through an angle $2\theta_1$ to circle Γ_1'. The circle Γ_1'' is the image of circle Γ_1' through the obstacle and the transmission coefficient is given by

$$t_{\text{db}} = (b'H)_{\text{db}} - (R)_{\text{db}}/2, \qquad (12)$$

where $(x)_{\text{db}}$ means $-20 \log_{10} |x|$ and R is the radius of circle Γ_1'' (see reference 5).

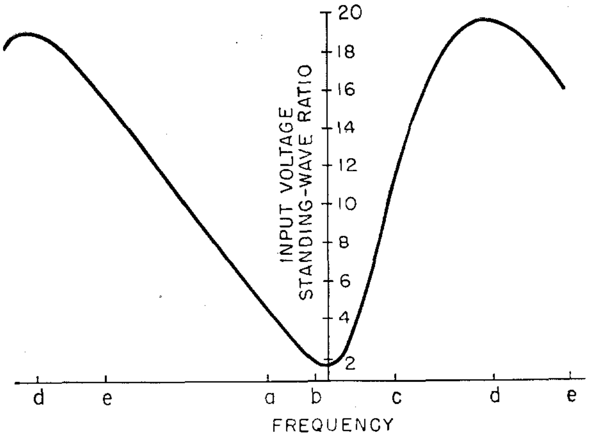

Figure 12—Input voltage standing-wave ratio plotted against frequency for a resonant section for lossy nonreversible obstacles.

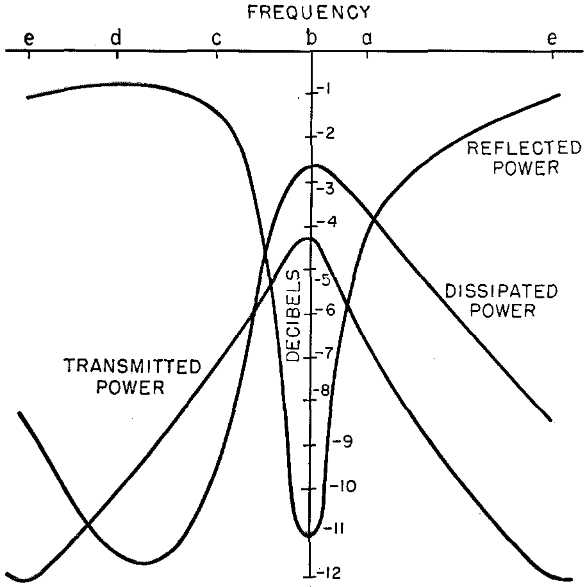

Figure 13—Relative reflected, transmitted, and dissipated powers in a resonant section for lossy nonreversible obstacles.

The input voltage standing-wave ratio, the transmitted and reflected power, and the dissipated power, deduced from the diagram of Figure 10 are shown in Figures 12 and 13. The curve for the transmitted power is symmetrical as should be expected from (10). The asymmetry of the curves of input voltage standing-wave ratio and reflected power versus frequency should be noted. It could be more or less pronounced depending on the particular values of the scattering-matrix coefficients of the obstacles.

D. As a practical example, the case of posts transverse to the microstrip will be considered.

The posts are 0.05 inch (1.3 millimeters) in diameter and located in the middle of a strip conductor 0.24 inch (6.1 millimeters) wide printed on *G-6* Fiberglas stock $\frac{1}{16}$ inch (1.6 millimeters) thick. At a frequency of 4900 megacycles per second and with the reference planes taken at the axis of the post, the scattering-matrix coefficients of the post as deduced from the circle diagram Γ_1 of Figure 14 are

$$\begin{aligned} S_{11} &= -\ 0.80 + j0.52 \\ \pm\, S_{12} &= \quad 0.17 + j0.23 \\ S_{22} &= -\ 0.77 + j0.55. \end{aligned}$$

The locus of the input reflection coefficient versus frequency of 2 such posts (located in the line and 14 millimeters apart, the line being terminated in a matched load) is the circle Γ_3. The experimental results shown in Figure 15 are in good agreement with the computed values shown in Figure 14. Similarly, the insertion loss at the center frequency has been found experimentally to be about 1.5 decibels and is also in agreement with the computed value.

In Figure 16, the input voltage standing-wave ratio has been plotted against frequency together with a similar curve in the case of lossless susceptances $|b| = 6.5$. The increase in bandwidth due to the losses introduced by the obstacles and by the line is noticeable. The double-loaded Q_L in this case is of the order of 26 and the unloaded Q_0 is equal to 164.

3.1 Discussion

In the case of a lossless obstacle having a large value of susceptance, the coefficient S_{11} can be obtained more accurately by using a frequency-resonance method.

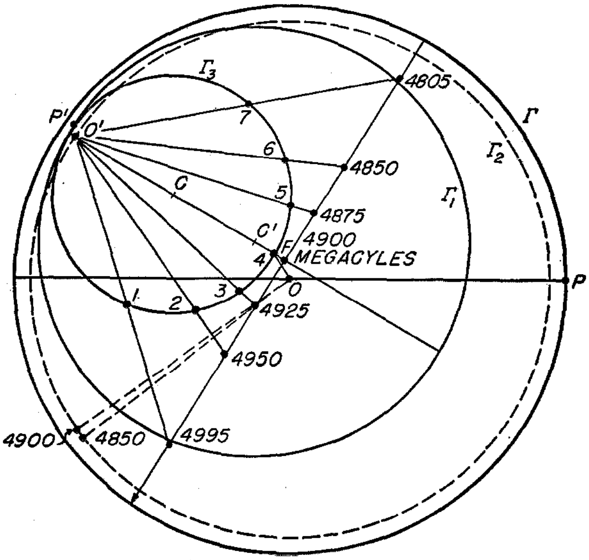

Figure 14—Frequency response and Q of a resonant section for lossy obstacles.

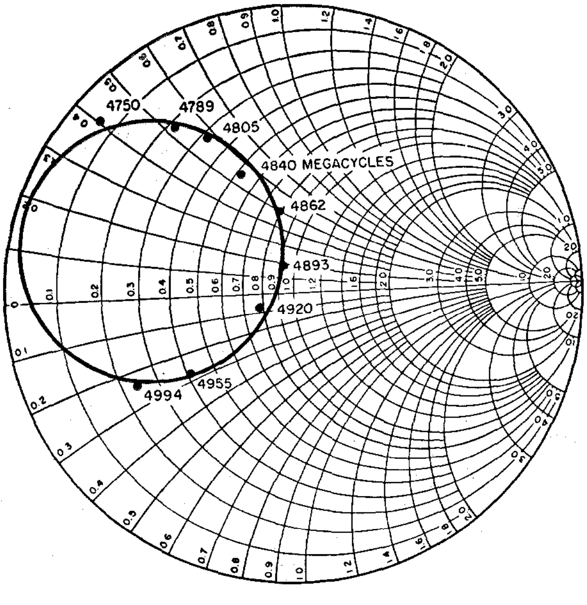

Figure 15—Frequency response and Q of a resonant section obtained experimentally.

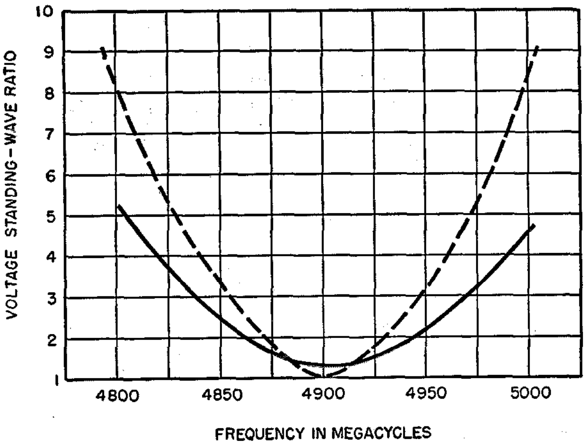

Figure 16—Input voltage standing-wave ratio plotted against frequency for a resonant section.

If a resonant section is made with two identical obstacles separated by a length of line L, from the relation

$$\Delta f = \frac{4}{\pi}\frac{f_0}{b^2} \qquad (7)$$

one can deduce easily

$$|S_{11}| = 1 \Big/ \left(\pi \frac{\Delta f}{f_0} + 1\right)^{1/2}, \qquad (13)$$

where Δf is the 3-decibel bandwidth and f_0 is the frequency at resonance.

A somewhat similar relation is obtained in the case of a lossy, nonreversible obstacle. For large values of susceptances ($|b| \geqslant 4$, for instance) the projection of points 1, 2, 3, . . . on an axis perpendicular to the line CO' defines a linear scale of frequencies and it has been shown by Deschamps that to a good approximation the distance $O'F$ is such that (see Figure 17)

$$O'F = 2R[A/(A+1)], \qquad (14)$$

where $2R$ is the diameter of circle Γ_3 and A is given by the relation

$$A = \frac{\lambda_0}{L}\frac{f_0}{\pi \Delta f}, \qquad (15)$$

λ_0 being the wavelength in microstrip at frequency f_0 and Δf, the 3-decibel bandwidth.

Usually the position of the point F is well defined experimentally. Then it is easy to obtain $OO' = S_{11}$ from the triangle $OO'F$, where OF and the angle β are obtained experimentally and $O'F$ is given by the above relation. This is the frequency-resonance method previously mentioned in section 2.2.

4. Conclusions

The Deschamps method is particularly well adapted to microstrip, since it gives directly the most-important parameters of the line or of obstacles in the line, independent of any equivalent circuit representation and arbitrary definition for the characteristic impedance of the line.

The examples discussed here should not be considered as the optimum performances that can be obtained with microstrip. Lines with lower insertion loss and obstacles having larger susceptances and less radiation loss have been developed already and it is hoped to give the results of some practical application in the near future.

5. Acknowledgment

The authors acknowledge the many original contributions of G. A. Deschamps during all phases of this work.

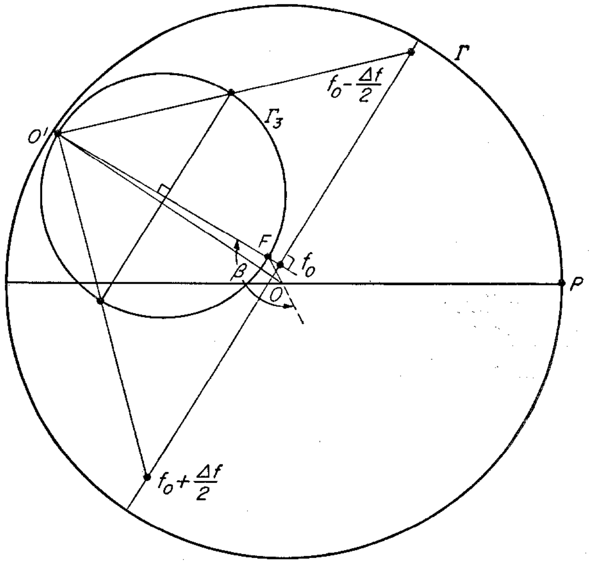

Figure 17—Frequency-resonant method for accurate determination of S_{11}.

6. Appendix

6.1 Application of Deschamps' Method to Microstrip

Recently, a method has been described[5] for the determination of the reflection coefficient and insertion loss of a junction in closed waveguides. Essentially, in this method the length of the line after the junction is varied over half a wavelength by moving a short-circuit and the corresponding reflection coefficient is measured at the input of the junction in the standard manner. The successive positions of the short-circuit are arranged to differ by $\lambda/8$. When plotted on the complex plane, these reflection coefficients fall on a circle, assuming the losses in the line to be small within half a wavelength. The chords that join opposite points intersect at a point that is related[5] in a simple manner to the reflection and transmission coefficient of the junction.

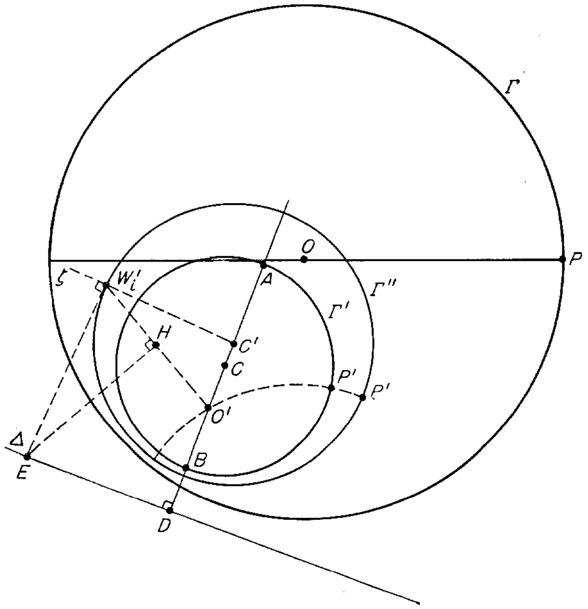

Figure 18—Γ'', image of a unit circle Γ through the junction. Γ'' is deduced from the circle Γ' of a lossy load, point W_i' corresponding to a lossless short-circuit, and the iconocenter O'.

This method has been adapted to microstrip. The specimen is connected through a junction to an ordinary coaxial or waveguide standing-wave machine. A good short-circuit is provided at the end of the line by soldering a large conductor plate between the strip conductor and ground plane.

This method, however, destroys the microstrip sample and there is no possibility of checking the measurements later. The following suggestion was made by Deschamps and the method thus modified has proved to be very practicable for the measurement of characteristics in microstrip.

It is possible to obtain the circle diagram Γ'' corresponding to the displacement of a good short-circuit without having to cut the end of the line continuously. To this effect and with reference to Figure 18, the reflection coefficient W_i', given by a good short-circuit for one position in the line, is compared with the circle diagram Γ', obtained by moving a given reflective load around this position. Knowing the iconocenter O' of circle Γ' and point W_i', the circle Γ'' can be constructed easily.

A constant reflective load can be obtained with a microstrip stub in shunt across the main line. Behind this obstacle, a matched load is provided so that no microwave energy that may pass through the obstacle is reflected back (see Figure 19). The obstacle is made of a copper foil secured to a piece of polyethylene tape. The obstacle is simply pressed against the strip conductor and maintained in place by the tape. The whole assembly is easy to move and to locate accurately.

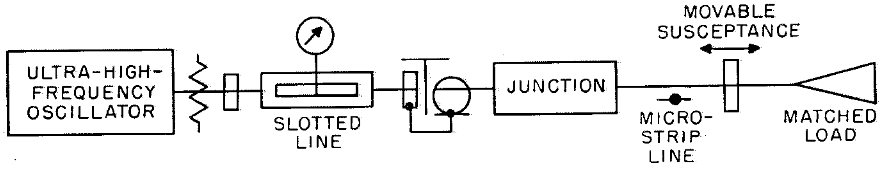

Figure 19—Experimental setup.

6.2 Note on Figure 18

For Figure 18, the measured quantities are W_i', the reflectance measured for one position of a good short-circuit in the line; Γ', the circle diagram of center C obtained by moving a reflective load around this position; and O', the iconocenter or image of the matched load.

The problem is to construct a circle Γ'' having a center C' and corresponding to the continuous displacement of a good short-circuit in the line. There are three steps in the solution.

A. Draw line Δ, a radical axis, perpendicular to line $O'C$ at D.

$$\overline{DO}'^2 = DB \times DA.$$

B. Obtain point E at intersection of Δ and the EH perpendicular bisector of $O'W_i'$.

C. Draw line ζ orthogonal to EW_i' at W_i'. The center C' of circle Γ'' is at the intersection of $O'C$ and ζ.

This construction is based on the properties that the radical axis Δ is the locus of the centers of a system of circles each orthogonal to all member circles of the family of circles defined by Γ' and O'.

Coaxial Line with Helical Inner Conductor*

By WILLIAM SICHAK

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TO OBTAIN an approximate solution for the fields in a coaxial line with a helical inner conductor, the helix is replaced by a fictitious surface that is conducting only in the helix direction, an approximation used in the early work on traveling-wave tubes. Maxwell's equations are solved for the lowest "mode" (all fields independent of angle) when the medium inside the helix has permittivity and permeability different from that of the medium surrounding the helix. Equations for the velocity along the axis, characteristic impedance, attenuation constant, and Q are given.

The significant parameter is $(2\pi Na)(2\pi a/\lambda)$, where N = number of turns per unit length, a = helix radius, and λ = wavelength. When this parameter is considerably less than 1, the velocity and characteristic impedance depend only on the dimensions. The dielectric inside the helix has only a second-order effect, while the dielectric outside the helix has a first-order effect. The wave appears to propagate along the helix wire with the velocity of light only when the outer conductor is very close to the helix; as the outer-conductor diameter is increased, the apparent velocity along the wire gradually increases and reaches a limiting value when the outer conductor is infinitely large. For the shapes generally used, the apparent velocity along the wire is rarely more than 30-percent greater than the velocity of light, but with an infinitely large outer conductor this velocity can be 2 or 3 times the velocity of light.

When $(2\pi Na)(2\pi a/\lambda)$ is greater than 1, the wave appears to propagate along the helix wire with the velocity of light, and the characteristic impedance depends only on the ratio of wavelength to helix radius. Introducing higher-dielectric-constant material inside or outside of the helix has a first-order effect, and the effect is the same whether the material is inside or outside the helix.

The outer-conductor loss is appreciable, about one-fourth to one-half the helix loss for the usual shapes with low velocities along the axis. The unloaded Q's can be about the same as with conventional coaxial lines. The Q does not depend on the length of the resonator if $(2\pi Na)(2\pi a/\lambda)$ is less than about 0.5, so that for frequencies below about 500 megacycles per second, the volume is considerably less than the volume of a conventional coaxial-line resonator. A few measurements to check the formulas for Q are presented.

• • •

Coaxial lines with helical inner conductors are used in many applications—in traveling-wave tubes, as delay lines, high-Q resonators, and high-characteristic-impedance transmission lines, and in extending microwave impedance-matching techniques to frequencies as low as 300 kilocycles. An analysis of this transmission line is given in this paper. The equations reduce to those previously published when *A*) the number of turns per unit length goes to zero (the standard coaxial line) and *B*) the outer conductor is removed (the helix used in traveling-wave tubes). The assumption that the electromagnetic wave travels along the helix wire with a velocity very close to that of light is not true except in extreme cases so that formulas based on this assumption may be in error.

Maxwell's equations are solved by replacing the helix with a fictitious surface that is conducting only in the helix direction. This method has been successfully used in the early work on traveling-wave tubes.[1,2] All fields are assumed to be independent of angle (the lowest "mode"). These assumptions have been questioned on

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* Reprinted from *Proceedings of the IRE,* volume 42, pages 1315–1319; August, 1954.

[1] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Company, Incorporated, New York, New York; 1950: page 229. See also J. R. Pierce, "Theory of Beam-Type Traveling-Wave Tubes," *Proceedings of the IRE,* volume 35, pages 111–123; February, 1947.

[2] L. L. Chu and J. D. Jackson, "Field Theory of Traveling-Wave Tubes," *Proceedings of the IRE,* volume 36, pages 853–863; July, 1948.

sound theoretical grounds, but the results given here are good enough for most engineering applications. The details of the derivation are given in the appendix.

After this paper was written, a translation of a paper[3] originally published in Russian was obtained. It presents a general solution of the problem, emphasizing the high-frequency behavior with an electron beam. There is no discussion or derivation of characteristic impedance, velocity, or Q, the main points of this paper. Another paper[4] gives an expression for the velocity and an approximate expression for the characteristic impedance.

1. *Velocity*

When the dielectrics on both sides of the helix are identical, the velocity along the axis is given by

$$(c/V)^2 = 1 + (M\lambda/2\pi a)^2, \qquad (1)$$

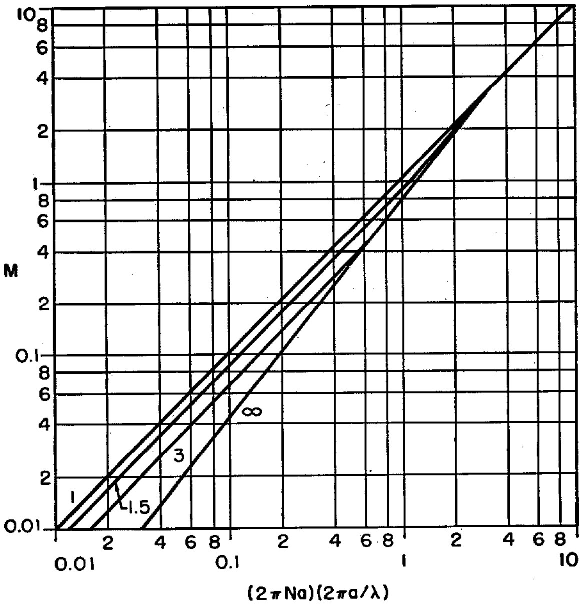

Figure 1—M plotted against $(2\pi Na)$ $(2\pi a/\lambda)$ for the indicated values of b/a.

$$(2\pi Na)\,\frac{2\pi a}{\lambda} = M\,\frac{J_0(jM)}{J_1(jM)}\left[\frac{\dfrac{H_0^{(1)}(jMb/a)}{J_0(jMb/a)} - \dfrac{H_0^{(1)}(jM)}{J_0(jM)}}{\dfrac{H_1^{(1)}(jMb/a)}{J_1(jMb/a)} - \dfrac{H_1^{(1)}(jM)}{J_1(jM)}}\right]^{1/2}. \qquad (2)$$

All symbols are defined in section 5.

These equations can be solved by plotting $(2\pi Na)(2\pi a/\lambda)$ against M, with b/a as a parameter as shown in Figure 1 for $\epsilon_1 = \epsilon_2$. Then if N, a, and λ are known, the velocity can be obtained from the first equation.

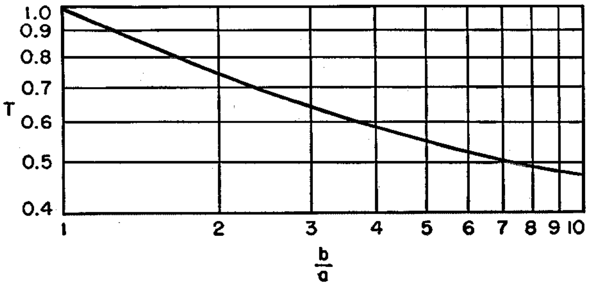

Figure 2—T in (3) is a function of b/a.

When the diameters are small $(b/\lambda < V/2\pi c)$, the above equations simplify to

$$\left.\begin{aligned}\left(\frac{c}{V}\right)^2 &\approx 1 + \frac{[1-(a/b)^2](2\pi Na)^2}{2\ln(b/a)}\\ &= 1 + T^2(2\pi Na)^2.\end{aligned}\right\}(3)$$

T versus b/a is plotted in Figure 2. Usually $(2\pi Na)^2$ is large, so that

$$\left.\begin{aligned}c/V &\approx T(2\pi Na)\\ &\approx (2\pi Na)(a/b)^{1/2}.\end{aligned}\right\}(4)$$

The last approximation becomes inaccurate for large values of b/a. The factor T is, for $(2\pi Na) \gg 1$, the ratio of the velocity of light to the velocity of the wave along the wire. This ratio does not depart markedly from 1 for most practical cases, but in the case of helical antennas[5] this ratio can be larger than 2.

When the diameters are large $(b/\lambda > V/c)$, the equations reduce to

$$(c/V)^2 \approx 1 + (2\pi Na)^2 = 1/\sin^2\psi. \qquad (5)$$

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[3] L. N. Loshakov, "Propagation of Waves along a Coaxial Spiral Line in the Presence of an Electron Beam," *Journal of Technical Physics*, volume 19, pages 578–595; May, 1949.

[4] C. O. Lund, "Broadband Transition from Coaxial Line to Helix," *RCA Review*, volume 11, pages 133–142; March, 1950.

[5] A. G. Kandoian and W. Sichak, "Wide-Frequency-Range Tuned Helical Antennas and Circuits," *Electrical Communication*, volume 30, pages 294–299; December, 1953: also, *Convention Record of the 1953 IRE National Convention*, volume 1, pages 42–45.

For this case (usually encountered in traveling-wave tubes), the wave travels along the wire with the velocity of light. When the diameters are small and the dielectric inside the helix is different from the dielectric outside the helix, the velocity is very nearly the same as if the dielectric were uniform throughout. On the other hand, for large diameters,

$$\frac{c_2}{V} \approx (2\pi Na)\left(\frac{\epsilon_1+\epsilon_2}{2}\right)^{1/2}. \tag{6}$$

This case is interesting because the velocity is the same whether the dielectric is inside the helix or outside of it and, in addition, does not depend on the ratio of diameters.

2. *Characteristic Impedance*

The voltage is the integral of E_{r2} from $r = a$ to $r = b$. The current is given by

$$I = 2\pi a H_{\theta 2}(r = a) - j\omega\epsilon_1 \int_0^a E_{x1}(2\pi r)dr,$$

$$Z_0 = \frac{c}{V} 30\pi J_0^2(jM)\left|\frac{H_0^{(1)}(jMb/a)}{J_0(jMb/a)} - \frac{H_0^{(1)}(jM)}{J_0(jM)}\right|. \tag{7}$$

Figure 3 shows how the characteristic impedance varies with M and b/a for $\epsilon_1 = \epsilon_2$.

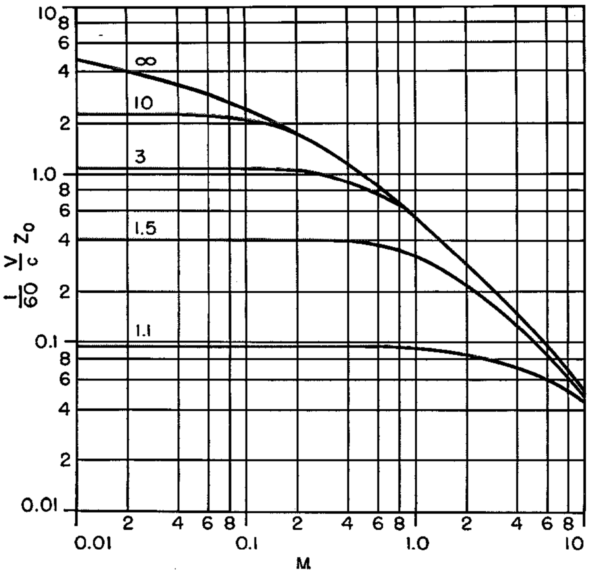

Figure 3—Relation of M to $(1/60)\ (V/c)Z_0$ for indicated values of b/a.

When the diameters are small, this equation reduces to

$$Z_0 \approx \frac{c_2}{V}\left(\frac{\epsilon_0}{\epsilon_2}\right)^{1/2} 60 \ln\left(\frac{b}{a}\right). \tag{8}$$

Using (3), this can be written

$$\left.\begin{aligned} Z_0 &\approx 120\pi Na\left[\frac{(1-a^2/b^2)\ln(b/a)}{2}\right]^{1/2}\left(\frac{\epsilon_0}{\epsilon_2}\right)^{1/2} \\ &= 120\pi NaT\,(\epsilon_0/\epsilon_2)^{1/2}\ln(b/a). \end{aligned}\right\} \tag{9}$$

As Winkler[6] has shown, the maximum value when holding N and a constant is obtained with $b/a = 2.06$.

This approximation holds when the curves in Figure 3 have zero slope. For the usual values of b/a (1.5 or greater), M must be less than about 0.4.

For $b/a = \infty$ and $\epsilon_1 = \epsilon_2$, the characteristic impedance is

$$Z_0 = (c/V)30\pi J_0(jM)H_0^{(1)}(jM). \tag{10}$$

For $M < 0.5$,

$$Z_0 \approx 60\frac{c}{V}\ln\frac{1.12}{M}. \tag{11}$$

For $M > 0.5$,

$$Z_0 \approx \frac{c}{V}\frac{30}{M} \approx \frac{30\lambda}{2\pi a}. \tag{12}$$

The last approximation holds when $c/V \gg 1$.

The standard formula for the inductance of a long solenoid can be obtained by treating the solenoid as a short length of short-circuited line and using (10) and (29).

3. *Losses and Q*

3.1 OUTER-CONDUCTOR LOSSES

The power loss in the outer conductor can be calculated since the tangential magnetic field is known.[7] The power lost per square meter of surface is

$$P_L = \pi b(H_x^2 + H_\theta^2)Z_{\text{wall}}. \tag{13}$$

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[6] M. R. Winkler, Discussion on "High Impedance Cable," by H. E. Kallman, *Proceedings of the IRE,* volume 35, page 1097; October, 1947.

[7] R. I. Sarbacher and W. A. Edson, "Hyper- and Ultra-High Frequency Engineering," John Wiley and Sons, Incorporated, New York, New York; 1943: page 262.

The attenuation constant is

$$\alpha = \frac{P_L}{2P_{\text{inc}}} = \frac{P_L}{2I^2 Z_0}. \tag{14}$$

For small diameters and high velocity ratios (the usual case of interest),

$$H_x \gg H_\theta,$$

$$I \approx 2\pi a H_{\theta 2}(r = a), \tag{15}$$

$$\alpha \approx \frac{Z_{\text{wall}}}{4\pi b Z_0}\left(\frac{a}{b}\right)^2 (2\pi N a)^2. \tag{16}$$

A result similar to (16) has been obtained by Bogle.[8] The factor $Z_{\text{wall}}/(4\pi b Z_0)$ is the attenuation constant due to the outer conductor in an ordinary coaxial line.

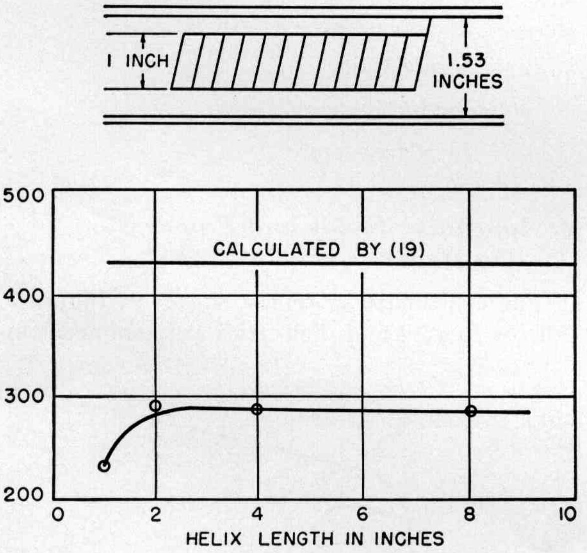

Figure 4—Experimental and calculated values of Q versus helix length. In the experimental helix, $Nd = 1/3$ and there were 116 turns with one end connected to the outer conductor.

3.2 HELIX LOSSES

The procedure used to calculate the outer-conductor losses leads to erroneous results when applied to the helix because the boundary conditions of the helix are not satisfied exactly. An approximate result can be obtained by assuming that the resistance of the helix is the same as an isolated conductor of the same diameter.

$$\alpha \approx \frac{Z_{\text{wall}}(2\pi N a)}{4\pi d Z_0} = \frac{Z_{\text{wall}}(2\pi N a)^2}{8\pi^2 a Z_0 (Nd)}. \tag{17}$$

The factor $(2\pi Na)$ is introduced because the unit of length is taken along the axis rather than along the helix wire.

This approximation does not take into account the proximity effect (apparent increase of resistance in a conductor due to the proximity of other conductors).

3.3 THE Q

The unloaded Q using copper conductors is given by[9]

$$Q = \beta/2\alpha. \tag{18}$$

For small diameters, this becomes

$$Q \approx \frac{120\pi a(1 - a^2/b^2) f_{mc}^{1/2}}{\dfrac{1}{2\pi N d} + \left(\dfrac{a}{b}\right)^3}. \tag{19}$$

This equation indicates that the optimum value of (Nd) is one as large as possible due to the approximations used in deriving these equations. However, it is known from experimental data that the optimum value of (Nd) is between 0.3 and 0.4.

The optimum value of b/a depends on (Nd). Using $(Nd) = 0.35$ and $b/a = 2.23$, the Q is

$$Q_{\text{opt}} \approx 250 b f_{mc}^{1/2}. \tag{20}$$

For a coaxial line with $b/a = 3.6$

$$Q_{\text{opt}} = 210 b f_{mc}^{1/2}. \tag{21}$$

These optimum Q's have been derived without regard to the resulting volume, subject only to the restriction that $(2\pi Na)(2\pi a/\lambda)$ be less than 0.5–1.0. The quantity *Q/volume* is a maximum for $(Nd) = 0.35$ when $b/a = 1.57$.

$$Q_{\text{opt}} \approx 200 b f_{mc}^{1/2}. \tag{22}$$

4. *Experimental Results*

According to (19), the Q does not depend on the length or volume of the resonator. Measurements were made on a series of resonators with the same helix and outer-conductor diameters but with different wire diameters and turns per inch in the helix, but holding the product (Nd) constant. The coils ranged in length from 1 inch to 8 inches (2.5 to 20 centimeters). The resonant frequencies were near 9 megacycles. Q versus helix length is shown in Figure 4. Over most of the range, the Q is constant and equal to about two-thirds of the calculated values. Some of the discrepancy between measurement and theory is due to the use of commercial copper, to the method of soldering the helix to the outer conductor, and to the use of a solid polystyrene form for winding the helices. The shortest helix has a characteristic impedance less than the others (because $(2\pi Na)(2\pi a/\lambda)$ is about 1), so that its Q should be smaller.

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[8] A. G. Bogle, "Effective Inductance and Resistance of Screened Coils," *Journal of the Institution of Electrical Engineers*, volume 87, pages 299–316; September, 1940.

[9] "Reference Data for Radio Engineers," Federal Telephone and Radio Company, New York, New York, third edition; 1949: pages 304–319.

5. Glossary

a = mean radius of helix

b = outer-conductor radius

$$B = \frac{p_1 \tan \psi J_0(p_1 a)}{j\omega\mu_1 J_1(p_1 a)}$$

c = velocity of light

d = radius of helix wire

$$D = \frac{J_0(p_1 a)}{-GJ_0(p_2 a) + N_0(p_2 a)}$$

E_x,
E_r,
E_θ = electric field in the indicated direction

f = frequency

f_{mc} = frequency in megacycles

$$F = \frac{p_2 \tan \psi J_0(p_1 a)}{j\omega\mu_2[-HJ_1(p_2 a) + N_1(p_2 a)]}$$

$$G = \frac{N_0(p_2 b)}{J_0(p_2 b)}$$

$$H = \frac{N_1(p_2 b)}{J_1(p_2 b)}$$

H_x,
H_r,
H_θ = magnetic field in the indicated direction

$H_0^{(1)}$,
$H_1^{(1)}$ = Hankel functions of the first kind

I = current

J_0,
J_1 = Bessel functions of the first kind

$M = |pa|$

N = number of turns per unit length

N_0,
N_1 = Bessel functions of the second kind

$p^2 = \gamma^2 + \omega^2 \epsilon\mu$

P_{inc} = incident power

Q = quality factor

V = velocity along the axis

Z_0 = characteristic impedance

$Z_{\text{wall}} = (\pi f \mu/\sigma)^{1/2}$

α = attenuation constant

$\beta = 2\pi/\lambda_a$ = phase constant along the axis

$\beta_0 = 2\pi/\lambda$ = free-space phase constant

$\gamma = \alpha + j\beta$ = propagation constant

ϵ = permittivity

λ = free-space wavelength

μ = permeability

σ = conductivity

$\omega = 2\pi \times$ frequency

6. Appendix—Fields and Propagation Constants

The coordinate system is shown in Figure 5. All conductors and dielectrics are assumed lossless and the fields are independent of θ. Inside the helix, the longitudinal fields are taken to be of the form

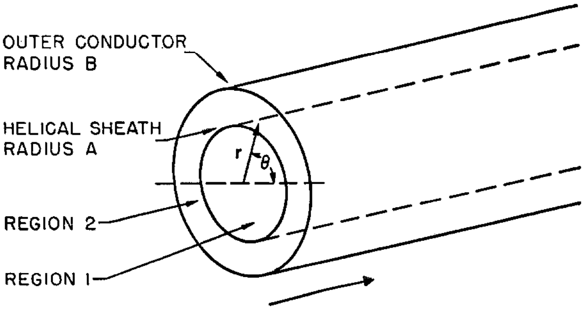

Figure 5—Coordinate system. ψ is the angle between a turn of the helix and the cross-sectional plane of the helix.

$$\left.\begin{aligned} E_{x1} &= J_0(p_1 r), \\ H_{x1} &= BJ_0(p_1 r). \end{aligned}\right\} \quad (23)$$

Bessel functions of the second kind are not used because they go to infinity at $r = 0$.

Between the helix and the outer conductor, the longitudinal fields are taken to be of the form

$$\left.\begin{aligned} E_{x2} &= D[-GJ_0(p_2r) + N_0(p_2r)], \\ H_{x2} &= F[-HJ_0(p_2r) + N_0(p_2r)]. \end{aligned}\right\} \quad (24)$$

The boundary conditions $E_{x2} = 0$ and $\partial H_{x2}/\partial r = 0$ at $r = b$ have been applied in the above equations.

By using Maxwell's equations[10] the remaining fields can be determined.

Inside the helix ($r \leqslant a$),

$$\left.\begin{aligned} E_{r1} &= \frac{\gamma_1}{p_1} J_1(p_1r), \\ E_{\theta 1} &\doteq \frac{(-)Bj\omega\mu_1}{p_1} J_1(p_1r), \\ H_{r1} &= \frac{B\gamma_1}{p_1} J_1(p_1r), \\ H_{\theta 1} &= \frac{j\omega\epsilon_1}{p_1} J_1(p_1r). \end{aligned}\right\} \quad (25)$$

Outside the helix ($a \leqslant r \leqslant b$),

$$\left.\begin{aligned} E_{r2} &= \frac{D\gamma_2}{p_2}[-GJ_1(p_2r) + N_1(p_2r)], \\ E_{\theta 2} &= \frac{-Fj\omega\mu_2}{p_2}[-HJ_1(p_2r) + N_1(p_2r)], \\ H_{r2} &= \frac{F\gamma_2}{p_2}[-HJ_1(p_2r) + N_1(p_2r)], \\ H_{\theta 2} &= \frac{Dj\omega\epsilon_2}{p_2}[-GJ_1(p_2r) + N_1(p_2r)]. \end{aligned}\right\} \quad (26)$$

The factor $\exp[j(\omega t - \gamma x)]$ multiplying the right-hand sides of all of the above equations has been omitted.

The boundary conditions at the helical sheet ($r = a$) are

$$\left.\begin{aligned} E_{x1} &= E_{x2}, \\ E_{\theta 1} &= E_{\theta 2}, \\ \frac{E_{x1}}{E_{\theta 1}} &= \frac{E_{x2}}{E_{\theta 2}} = -\cot\psi, \\ (H_{x1} - H_{x2}) &+ (H_{\theta 1} - H_{\theta 2})\cot\psi = 0. \end{aligned}\right\} \quad (27)$$

For a nontrivial (that is, one in which the coefficients B, D, and F are not zero) solution to exist, the following equation must hold:

$$\begin{aligned} &\omega^2 \frac{\mu^2}{p^2}\cot^2\psi \\ &\times\left\{\frac{\epsilon_2}{p_2}\left[\frac{-GJ_1(p_2a)+N_1(p_2a)}{-GJ_0(p_2a)+N_0(p_2a)} - \frac{\epsilon_1 J_1(p_1a)}{p_1 J_0(p_1a)}\right]\right\} \\ &= -\left(\frac{\mu_2}{\mu_1}\right)\left(\frac{p_1}{p_2}\right)\frac{J_0(p_1a)}{J_1(p_1a)} \\ &\qquad + \left[\frac{-HJ_0(p_2a)+N_0(p_2a)}{-HJ_1(p_2a)+N_1(p_2a)}\right]. \end{aligned} \quad (28)$$

This equation determines the propagation constant when the dimensions, dielectric constants, et cetera, are specified. When the outer conductor is removed, the equation reduces to that given by Harris.[11] When the dielectric on both sides of the helix is identical and the outer conductor is removed, the equation becomes

$$\frac{\beta_0^2 \cot^2\psi}{p^2} = -\frac{J_0(pa)H_0^{(1)}(pa)}{J_1(pa)H_1^{(1)}(pa)}. \quad (29)$$

This result is the same as that given by Pierce,[1] and by Chu and Jackson.[2]

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[10] See page 243 of footnote reference 7.

[11] L. A. Harris, H. R. Johnson, A. Karp, and L. D. Smullin, "Some Measurements of Phase Velocity along a Helix with Dielectric Supports," Massachusetts Institute of Technology Radiation Laboratory Report 93: January 21, 1949.

United States Patents Issued to International Telephone and Telegraph System August-October, 1954

UNITED STATES patents numbering 31 were issued between August 1 and October 31, 1954 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that are of more-than-usual interest are included. In cases where corresponding Canadian patents have already issued, their numbers are given in parentheses.

H. H. Adelaar, System for the Generation of Electric Pulses, 2,689,910.

A. J. Baracket and E. Stein, Automatic Synchronization, 2,686,833 (644,716).

J. I. Bellamy, Impulse-Repeating Electromagnetic Relay, 2,689,883.

J. I. Bellamy and P. W. Hemminger, Receiver for Remote Supervisory and Control Systems, 2,688,738.

H. Bosse, Slot Antenna, 2,686,265.

H. Boyer, Block Coupler, 2,686,841.

W. C. Bruckman and M. W. Kretichman, Washing-Machine Support, 2,687,861.

M. P. G. Capelli, Radio Altimeter, 2,686,302 (609,614).

A. M. Casabona, Null-Type Glide-Slope System, 2,685,688 (655,752).

E. Christensen, Automatic Telecommunication System with Absent-Subscriber Service, 2,686,842.

G. Deakin, Traffic-Recording System for Automatic Telephone Exchange, 2,686,836.

M. den Hertog, Group-Selection Control Circuit, 2,686,839 (169,273).

M. den Hertog, Identification Circuit for Automatic or Semiautomatic Telephone Systems, 2,686,840.

M. den Hertog, C. de Zeeus, and H. H. Adelaar, Means for Automatically Detecting a Change of Condition in Any One of a Number of Electrical Circuits from a Common Circuit, 2,688,662 (610,228).

E. L. Earle, General-Purpose Relay, 2,686,850.

G. C. Hartley, F. H. Bray, and M. C. Branch, Register Translator for Telecommunication Switching Systems, 2,686,224 (597,150).

W. Hauer, Circuit Arrangement for Communication Systems, Particularly Private Automatic Telephone Branch Exchanges, 2,690,478.

G. H. Hough and T. M. Jackson, Gaseous Electric-Discharge Tube, 2,686,273 (624,908).

J. B. Lair, Impulse-Storing and Distributing Circuit, 2,691,727.

G. X. Lens, Sorting Machine for Letters or Similar Flat Objects, 2,689,657 (623,914).

M. R. Mauge, Switching System for Grouped Telephone Lines, 2,692,917.

A. J. Montchausse, Crossbar Multiselector Arrangement, 2,691,700 (630,286).

A. J. Montchausse, Multiswitch Apparatus Controlled by Crossbars, 2,686,226.

A. D. Odell, Electrical Switching Circuit, 2,688,695.

D. C. Rogers and W. W. Marsh, Thermionic Cathode, 2,686,272 (626,223).

L. E. Setzer and C. B. Watts, Jr., Automatic Approach Control, 2,689,345.

R. Urtel, Horizontal-Line Registration for Pickup Tubes, 2,691,743.

A. R. Vallarino and S. W. Lewinter, Radio Dispatching System, 2,685,642.

S. van Mierlo, Electronic Switching, 2,688,661.

E. P. G. Wright, D. A. Weir, and J. Rice, Means for Checking Recorded Information, 2,688,656 (608,264).

H. N. Wroe, Telephone Signaling Arrangement, 2,686,228 (627,411).

Radio Dispatching System[1]

A. R. Vallarino and S. W. Lewinter
2,685,642—August 3, 1954

This patent covers a radio system for dispatching and following the movement of vehicles such as taxicabs or police patrol cars in separate zoned areas. Only three frequencies are used and the areas are so positioned that no contiguous areas operate on the same frequency channel. A central dispatching station is used to transmit frequencies in radial sector-shaped zones. This central dispatching area may be then surrounded by other zones that do not have contiguous boundaries operating on the same frequency.

Gaseous Electric-Discharge Tube

G. H. Hough and T. M. Jackson
2,686,273—August 10, 1954

A gaseous discharge tube is described having several spaced cathodes and a common anode, the cathodes being in the form of metallic strips mounted in a common plane and having substantially equal spacing between adjacent strips. The anode is a metallic strip spaced from these cathodes and at right angles thereto. A shielding means is provided between pairs of cathodes to reduce the ionization coupling between the cathodes of each pair.

Sorting Machine for Letters or Similar Flat Objects[2]

G. X. Lens
2,689,657—September 21, 1954

A sorting device for letters is in use in the Belgian Post Office in Antwerp and other places. It consists essentially of an endless conveyor arrangement with carriers that move past receptacles for the letters, which are disposed in the carriers in vertical position. The letters will be properly dropped into the various receptacles in a substantially vertical direction.

Automatic Synchronization

A. J. Baracket and E. Stein
2,686,833—August 17, 1954

An automatic synchronization system for television monitoring use in which a control voltage is produced in response to the presence or absence of synchronizing pulses in the video picture source; there is also provided an independent source of synchronizing pulses together with a gating means to permit these independent pulses to pass if the synchronizing pulses are not present in the applied video signal.

Null-Type Glide-Slope System

A. M. Casabona
2,685,688—August 3, 1954

A null-type glide-slope system having an additional feature that tends to eliminate ambiguities that might be encountered in the glide slope. The original glide-slope array comprises two antennas, one spaced at a height H and the other at a height $H/2$ above ground plane and energized to produce different sidebands characteristic of the glide-slope signals. The ambiguities are removed by providing two other antennas energized 180 degrees out of phase with energy of the carrier frequency plus one sideband only and spaced above and below the lower of the other two antennas at heights of $2H/3$ and $H/3$, respectively.

Automatic Approach Control

L. E. Setzer and C. B. Watts, Jr.
2,689,345—September 14, 1954

An automatic approach-control arrangement for aircraft in which the usual controls for correcting the departure of the craft from the desired beacon course line are supplied and in addition thereto, there is derived a third signal indicative of the direction of departure off-course, but substantially independent of the degree of this departure. This third signal is combined in opposition to the guiding signals to provide the resultant control voltage. By this means the degree of correction is reduced as the proper course is approached so that course swings are reduced.

---

[1] A. R. Vallerino and S. W. Lewinter, "Radio Dispatching System for Operation of a Large Taxicab Fleet," *Electrical Communication,* volume 30, pages 55–60; March, 1953: also, *Electrical Engineering,* volume 71, pages 232–235; March, 1952.

[2] S. Scheuer, C. B. Neyt, and L. J. G. Nijs, "Electromechanical Distributor for Sorting Mail," *Electrical Communication,* volume 28, pages 163–170; September, 1951.

Contributors to This Issue

Maurice Arditi

Maurice Arditi was born on March 1, 1913, in Paris, France. In 1933, he received a degree in engineering physics at the Ecole de Physique et Chimie Industrielle of Paris, in 1934 a degree in electrical engineering at Ecole Supérieure d'Electricité, and in 1935 an M.S. degree in physics and chemistry from Sorbonne University, Paris.

From 1936 to 1939, Mr. Arditi did research work in physical chemistry at Paris University. In 1939, he joined the laboratories of Le Matériel Téléphonique, Paris, working on secondary emission, electron multipliers, and electron optics. He joined Federal Telecommunication Laboratories, New York, in 1944 and resumed work on special switching tubes. From 1947 to 1951, he was active in the field of microwave receivers for radio links and is presently engaged in the development of the microstrip line. Some of his studies on this subject are reported in the present issue.

Mr. Arditi is a Senior Member of the Institute of Radio Engineers.

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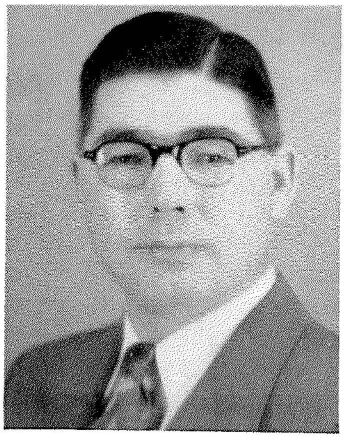

Sterling O. Bickel

Sterling O. Bickel was born in Union City, Indiana, on January 18, 1921. He served with the Army Engineer Corps and was released in 1947. In 1949, Mr. Bickel received B.S. degrees in radio and in administrative engineering from Tri-State College, Angola, Indiana.

Mr. Bickel was with the Magnavox Company until 1951, at which time he joined Capehart–Farnsworth Company as a tube-reliability engineer in guided missiles. Some of his work on subminiature-tube reliability is reported on in this issue.

Mr. Bickel is an Associate Member of the Institute of Radio Engineers.

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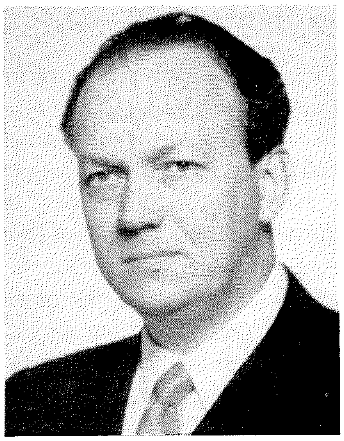


Amund Braaten

Amund Braaten was born on May 11, 1905 in Oslo, Norway. He studied electrical engineering at the Technical University of Norway, graduating in 1929.

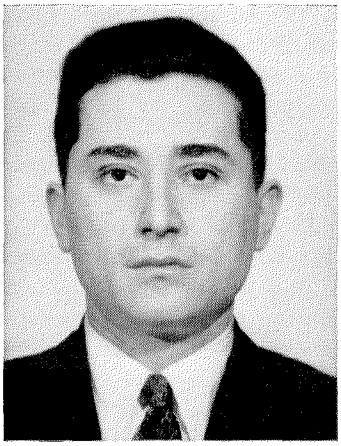
Following graduation, he joined the Standard laboratories in Hendon, England, returning to Norway in 1931, where for a brief time he served as a member of the staff at the Technical University.

Mr. Braaten rejoined the International System in 1932 as engineer of manufacture and cable engineer at Standard Telefon og Kabelfabrik A/S. He was promoted to chief engineer in 1941, to works manager in 1946, and to managing director in 1953. He reports in this issue on the new high-voltage cable-testing facilities recently installed at Oslo.

• • •

Leopold Christiansen was born in Kiev, Russia, on December 2, 1913. He studied at the Technical University of Stuttgart, at the Rensselaer Polytechnic Institute of Troy, New York, and at the Technical University of Dresden, where he received his diploma and his doctorate.

Leopold Christiansen



Jack Elefant

He worked first at the Institute of Electrical Engineering of the Technical University of Dresden. In 1948, he joined the Standard Elektrizitäts-Gesellschaft AG (division of Mix & Genest). Since that time, he has been engaged in the development of various carrier-frequency telephony systems.

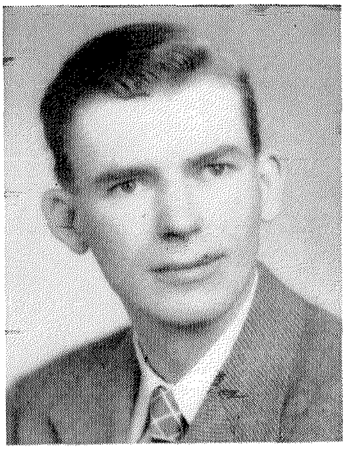
In this issue, he is author of a paper on distortion in carrier-frequency modulators.

• • •

Jack Elefant was born in New York City on October 8, 1923. He received a bachelor's degree in electrical engineering from the College of the City of New York in 1944 and a master's degree in 1951 from Polytechnic Institute of Brooklyn.

After several years with the Board of Transportation of New York City, he joined the staff of Federal Telecommunication Laboratories in 1951. A report on some of his work on microstrip lines is included in this issue of *Electrical Communication.*

Mr. Elefant is now associated with Airborne Instrument Laboratories of Mineola, New York.



Henry G. Nordlin

• • •

Henry G. Nordlin was born in Saint Paul, Minnesota, on September 1, 1921. He received the B.E.E. degree from the University of Minnesota in 1943 and has since completed his studies for an M.S. degree in electrical engineering at Stevens Institute of Technology.

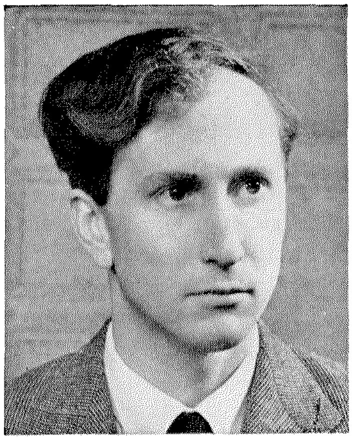
He joined the Federal Telephone and Radio Company in 1943 and is now a project engineer in the chemical and physical division of Federal Telecommunication Laboratories.

Mr. Nordlin is a Senior Member of the Institute of Radio Engineers.

• • •

D. S. Ridler was born in London, England, in 1921. He joined the telephone switching division of Standard Telephones and Cables, Limited, in 1939 and later transferred to Standard Telecommunication Laboratories, Limited, where he has been engaged in work on the design and application of digital devices.

Mr. Ridler is the coauthor of the paper in this issue on the application of functional diagrams to electronic telegraph circuits. He is an Associate Member of the Institution of Electrical Engineers.

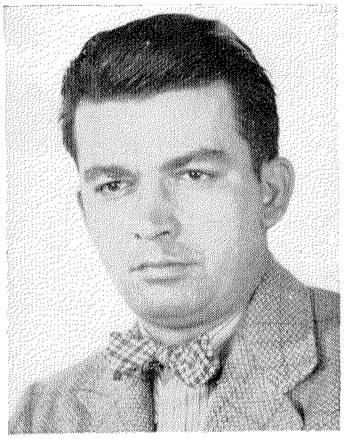


D. S. Ridler

William Sichak was born on January 7, 1916, in Lyndora, Pennsylvania. He received the B.A. degree in physics from Allegheny College in 1942.

From 1942 to 1945, he was engaged in developing microwave radar antennas at the Radiation Laboratory of Massachusetts Institute of Technology. Since then, he has been with Federal Telecommunication Laboratories, working on microwave antennas and allied equipment. In this issue, he is author of a paper on coaxial lines with helical inner conductors.

Mr. Sichak is in charge of the antennas and components division in the laboratories. He is a Member of the



William Sichak

Institute of Radio Engineers and of the American Physical Society.

• • •

E. P. G. **Wright**

E. P. G. **Wright** joined the equipment engineering branch of the Western Electric Company in London in 1920, and in 1926 took charge of project engineering when the company became associated with the International System. In 1928, Mr. Wright was transferred to the laboratories to work on switching problems. He was in charge of switching-system design for Standard Telephones and Cables, Limited, from 1932 until the war, when he was assigned to special duties on radio production and test-equipment design.

Mr. Wright is an author of the paper in this issue on functional diagrams for the design of telegraph circuits.

Mr. Wright served as a member of the British Telephone Technical Development Committee from its inception until 1939. He is a member of the Institution of Electrical Engineers and was awarded the Fahie Premium for papers on the bypath system and voice-frequency signalling and dialling.

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ELECTRICAL COMMUNICATION

Technical Journal of the
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and Associate Companies

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STABLE LABORATORY-STANDARD CAPACITOR MICROPHONE
MANUFACTURE OF TYPE-7400 TELEPHONE SELECTOR
CHARACTERISTICS AND APPLICATIONS OF THERMISTORS
GERMANIUM POSITIVE-GAP DIODE
BROAD-BAND REFLECTOMETERS AT HIGH FREQUENCIES
PULSE GENERATOR FOR MECHANOELECTRONIC SWITCHING SYSTEM
UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM

Volume 32 **JUNE, 1955** **Number 2**

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Volume 32 JUNE, 1955 Number 2

CONTENTS

Stable Laboratory–Standard Capacitor Microphone 75
By J. F. Houdek, Jr.

Manufacture of Type-7400 Step-by-Step Telephone Selector 82
By J. L. Donceel

Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors 98
By J. W. Howes

Germanium Positive-Gap Diode: New Tool for Pulse Techniques 112
By A. H. Reeves and R. B. W. Cooke

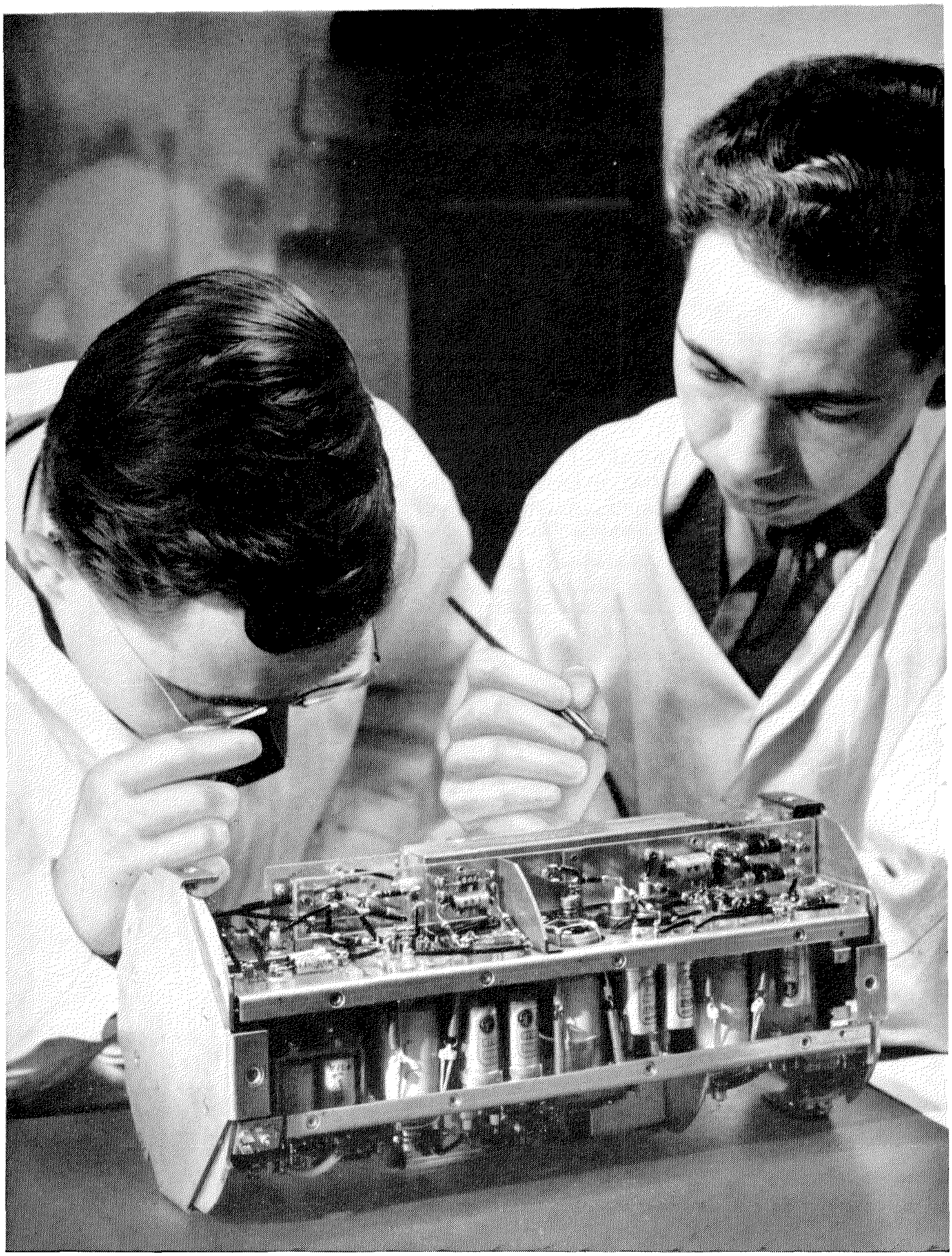
Broad-Band Reflectometers at High Frequencies 118
By R. T. Adams and Alexander Horvath

Pulse Generator for the Mechanoelectronic Telephone Switching System 126
By Eugene Van Dyck and Camille Weill

United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955 139

Recent Telecommunication Developments
- Back Issues of Electrical Communication 111
- Etching of E. H. Armstrong 138

Contributors to This Issue 141



The amplifier unit of a submerged repeater is shown undergoing inspection after assembly. Equipment similar to the above will be used in submerged repeaters designed jointly by the British Post Office and Standard Telephones and Cables, Limited for use in the Nova Scotia–Newfoundland portion of the forthcoming transatlantic telephone cables. Twenty-one of these repeaters are being manufactured by Standard Telephones and Cables in a special air-conditioned manufacturing unit. Each repeater will carry 60 two-way telephone circuits.

Stable Laboratory-Standard Capacitor Microphone*

By J. F. HOUDEK, JR.

Kellogg Switchboard and Supply Company, a division of International Telephone and Telegraph Corporation; Chicago, Illinois

LABORATORY-STANDARD microphones must provide dependable calibration stability, satisfactory performance, and reliability. Rugged construction together with a spring-loaded rear-electrode assembly permit the maintenance of constant diaphragm tension for improved stability. Appropriate modifications of the acoustic controls provide either a pressure-type or a free-field type of response. Each microphone has uniform characteristics and meets the American Standards Association performance requirements in its own applications. Supporting performance data and other information relative to capacitor microphones are presented.

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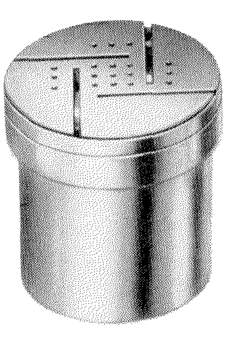

New capacitor microphone in actual size.

The microphone to be described does not differ in principle from the usual precision capacitor microphone in that it is essentially a stretched circular membrane near and parallel to a circular electrode. Since the theoretical aspects of these microphones and the effects of basic acoustic controls on their performance have been extensively treated[1,2] in the literature, this phase will not be covered here. Historically the development of capacitor microphones in the laboratory of the Kellogg Switchboard and Supply Company was started during 1932. In 1935, two practical designs were in use and as may be seen in Figure 1, they differed in size by about four to one. Both types were used for more than ten years before preference and development of the art caused the larger to lapse into obsolescence. Subsequently, only the smaller model of this microphone has been manufactured. Microphones of the early designs were incorporated in different types of test equipment where they gave satisfactory performance especially with respect to stability of calibration and adequacy of response–frequency characteristics. Design improvements of a minor nature have been made from time to time as a result of experience in the use and application of the microphone. Beginning with the second world war, Kellogg microphones were made available to other users, such as governmental laboratories, educational institutions, and manufacturing companies.

The present preferred method of calibration, the reciprocity method,[3,4] was not in general use when the microphones were first made. Level calibrations were made by the comparison method, using a calibrated microphone for reference. Response–frequency performance was determined by the electostatic-actuator method. After developing the reciprocity calibration apparatus shown in Figure 2, it was found that agreement between the two methods was within 0.5 decibel up to and including 13 000 cycles per second. Even after the reciprocity method became available, it was found to be advantageous to use the actuator to obtain relative sensitivities to frequencies extending well above the accurate range of the reciprocity apparatus, since calibrations by the actuator are not limited by frequency and are unaffected by the characteristics of the intervening gas.

---

* Reprinted from *Journal of the Audio Engineering Society*, volume 2, pages 234–238; October, 1954.

[1] P. M. Morse, "Vibration and Sound," 1st edition, McGraw-Hill Book Company, New York, New York; 1936: pages 160–173.

[2] T. R. Bonn, "Ultrasonic Condenser Microphone," *Journal of the Acoustical Society of America*, volume 18, pages 496–502; October 1946.

[3] L. L. Beranek, "Acoustic Measurements," 1st edition, John Wiley and Sons, New York, New York; 1949: pages 113–148.

[4] A. L. DiMattia and F. M. Wiener, "On the Absolute Pressure Calibration of Condenser Microphones by the Reciprocity Method," *Journal of the Acoustic Society of America*, volume 18, pages 341–344; October 1946.

More recently, emphasis was placed on free-field comparisons of two microphones, slightly different in contour, to determine and if necessary to eliminate any undesirable characteristics

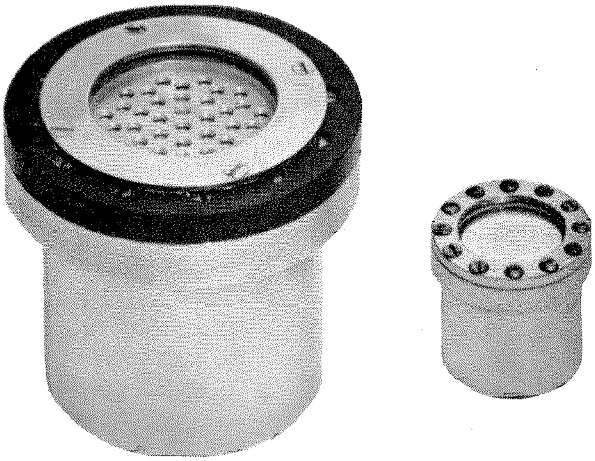

Figure 1—Two designs of microphones in use about 1935.

attributable to the size and shape of a microphone. The tests indicated no important undesirable characteristics and only a slightly greater directivity in the field pattern of the present free-field type over a proposed improved type.

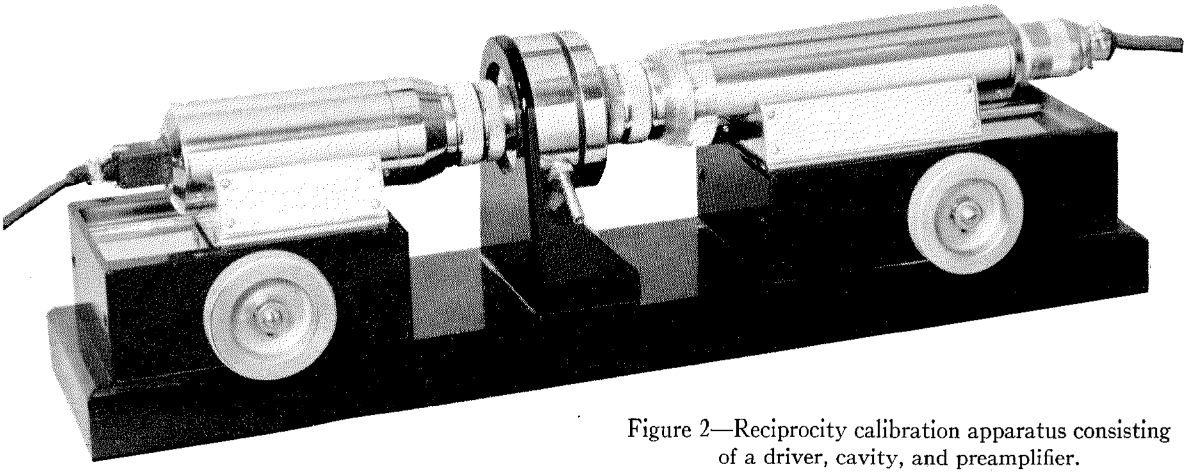

Figure 2—Reciprocity calibration apparatus consisting of a driver, cavity, and preamplifier.

1. Design Considerations

With regard to performance, all of the essential design considerations deemed necessary for the microphone were not specifically stated at the beginning of the development. With time and study, the following requirements were evolved.

A. The device should faithfully transduce speech and music at the intensities and frequencies normally encountered in audio work. Since it was to be the standard for the company in both the laboratory and the factory, special emphasis was placed on the ruggedness and stability of the unit.

B. The diaphragm size was restricted so as to be compatible with the preferred dimensions of the 6-cubic-centimeter artificial ears and couplers used in telephone receiver work.

C. From the diaphragm diameter so determined the design of the body was developed to be as small as possible, but with ample wall thickness to insure stability of performance. A flange was provided to simplify the mounting and positioning of the microphone with respect to the coupler or cavity with which it was to be used. This flange also added considerable rigidity to the body and further protected the diaphragm against possible strains due to mounting in couplers.

D. For free-field applications, such as the calibration of artificial voices, measurement of sound levels, studio and broadcast pickup, and the like, it was deemed advantageous to use the same microphone construction and to alter only the acoustic controls to attain more-uniform free-field characteristics.

E. In later stages of the development, commercial requirements led to higher insulation resistance, an extension of the range of relatively uniform frequency response, and to linearity at higher sound levels. The need for an unquestionably accurate method of calibration of the microphone led to the development of the reciprocity calibration apparatus. A minor refinement was required to prevent measurable hydrogen leakage when the microphone is used in its cavity.

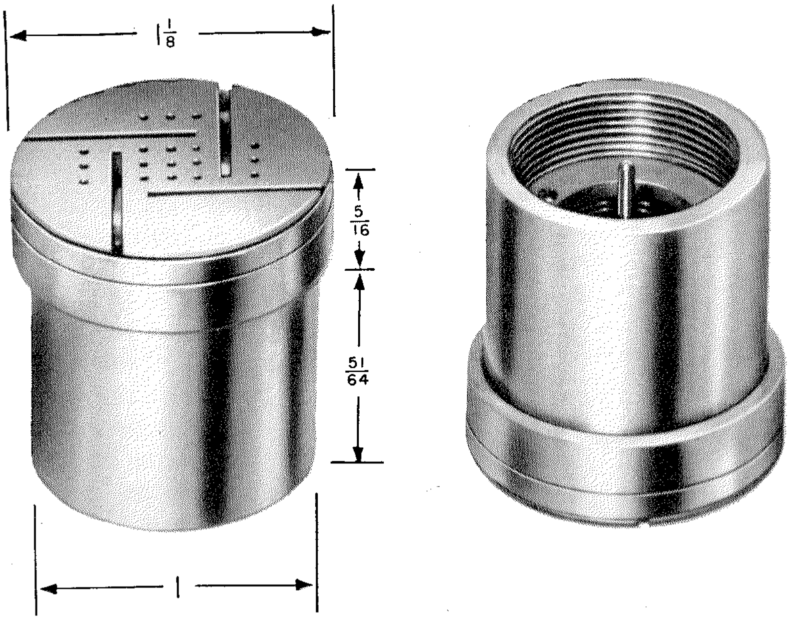

Figure 3—Dimensions in inches of the present model.

F. In the capacitor microphone, a compromise is involved between sensitivity and other performance factors. With emphasis on reliability and lowered distortion, the resulting sensitivity lies slightly below that of the highest-output microphones of this class.

2. *Design*

Externally, the microphone shown in Figure 3 has an over-all length of $1\frac{1}{8}$ inches (29 millimeters) without the grille. The main body diameter is 1 inch (25 millimeters) and the flange at the front end has a diameter of $1\frac{1}{8}$ inches (29 millimeters). The heavy wall thickness and the flange to which the diaphragm is clamped provide excellent dimensional stability. The microphone is furnished with a removable protective grille, which should not be used in couplers.

Figure 4 illustrates the internal construction and discloses the spring loading used to maintain constant tension on the diaphragm under slight dimensional changes (which might result from temperature variations and/or stretching of the diaphragm). The movable back-electrode assembly is shown well separated from the diaphragm to clarify constructional details. Also evident is the large contact area between the tensioning edge and the diaphragm, which serves as an air seal and obviates the necessity of placing the microphone in a desiccator between periods of use. The resonant cavity, although not clearly shown, is located behind the fixed electrode and is coupled to the cavity of 0.0008 to 0.0012 inch (0.02 to 0.03 millimeter) in height at the rear of the diaphragm by means of 8 intercommunicating holes. The reaction of

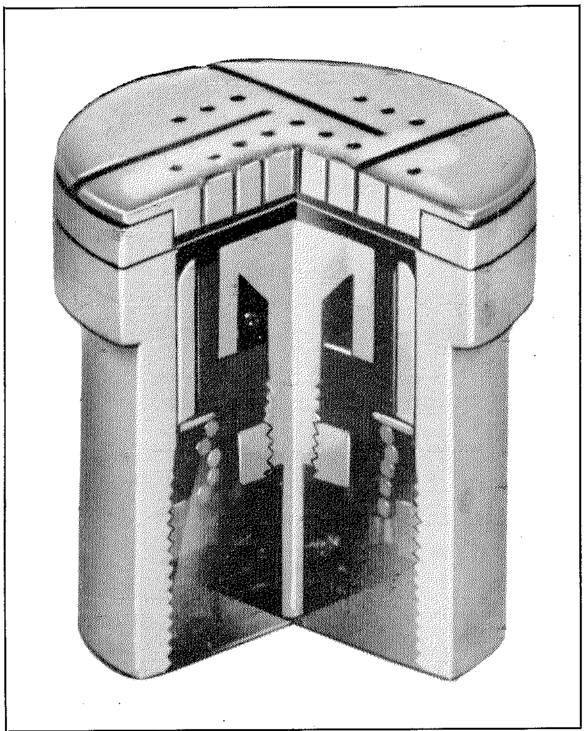

Figure 4—Sectional view of microphone.

this acoustical circuit has been carefully chosen to provide a compromise that flattens the resonance of the diaphragm and extends the frequency range of response.

The microphone has good shock resistance to all forces except those directed approximately perpendicular to the diaphragm. The weight is approximately 80 grams (2.8 ounces) and the unit is mounted by a substantial thread. Pressure should not be applied to the central electrode pin; a sliding-action spring-clip connection as shown in the adapter unit of Figure 5 is recommended. An adapter unit is used when it is necessary to couple the $\frac{25}{32}$-inch, 32-thread-per-inch fitting of the microphone to preamplifiers that have a $\frac{29}{32}$-inch, 60-thread-per-inch coupling.

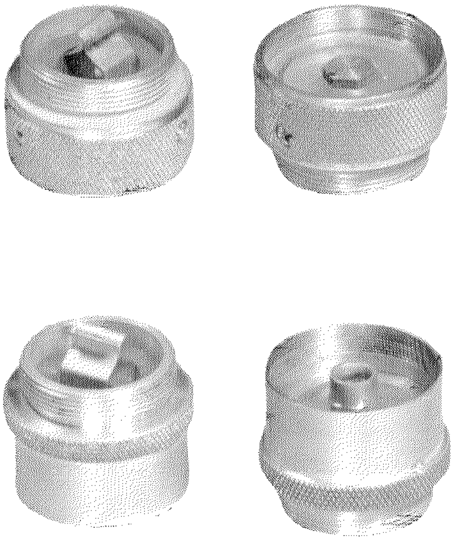

Figure 5—Adapter units.

3. *Performance Characteristics*

In studies on the capacitor microphone under conditions that were more severe than are encountered in normal use, it was found difficult in many cases to segregate the effects of environmental influences on the associated equipment. At other times, these factors cannot be extracted since the changes occurring in the microphone are of the same order as the accuracy of the testing equipment and it is necessary to express the results in terms of data obtained from the entire system. It is advantageous to attach the preamplifier directly to the microphone to obtain consistent and reproducible laboratory results.[5] Since the output impedance of the microphone is high in the audio-frequency range and requires a very-high load impedance, the usual cathode-follower is recommended for audio work. Various commercial preamplifiers are available and adaptable as an extension to the microphone, which then forms the nosepiece. At present in the laboratory, use is made of a modified Cruft cathode-follower-type circuit that is flat from about 1 to over 50 000 cycles.[6] For free-field work, both the above and smaller preamplifiers made by the Audio Instrument Company have been used.

Typical response–frequency data on type *A* and *B* microphones are summarized in Figures 6 and 7. Referring to Figure 6, the open-circuit voltage response as a function of the frequency of a constant sound source is shown for free-field and cavity conditions. The performance of the type-*A* microphone in cavities, such as that of the reciprocity coupler in an atmosphere of hydrogen, is shown as curve *C*. The response rise of the first resonant mode of vibration of the diaphragm is not removed entirely but by appropriate control, uniformity of response with respect to the low and middle frequencies as a base is improved.

However, if the cavity-type microphone is used in a free field, the remaining resonant-frequency rise plus the diffraction effects to sounds at perpendicular incidence to the microphone diaphragm result in the usual somewhat excessive peak in the 10-kilocycle region. This field pattern is, however, the usually accepted characteristic for a microphone designed for both cavity and free-field applications. Again with reference to curve *C*, it is clearly indicated that the microphone provides a uniform response in cavity work. Also included in this design is the consideration that higher sound intensities are obtained in couplers, especially in tests concerned with overload and high-intensity linearity tests on receivers. Therefore, the type-*A* cavity micro-

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[5] L. L. Beranek, footnote reference 3, pages 221–223.

[6] United States Office of Scientific Research and Development, "Report 3105 on Sound Pressure Meter"; 1944: pages 155–170.

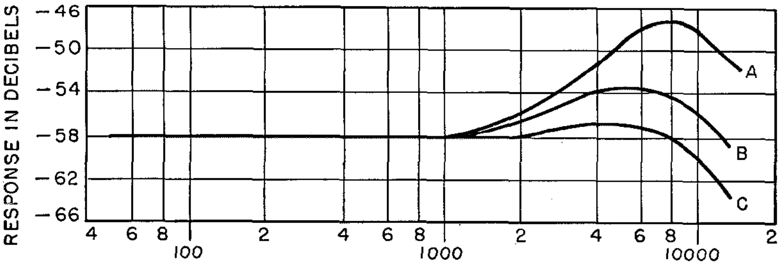

Figure 6—Voltage–frequency characteristic of type-*A* microphone. Open-circuit output voltage is plotted in decibels referred to 1 volt per dyne per square centimeter. *A* is for free-field operation with the grille and for sound incidence perpendicular to the diaphragm; *B* is for parallel incidence. *C* is for pressure response without the cavity.

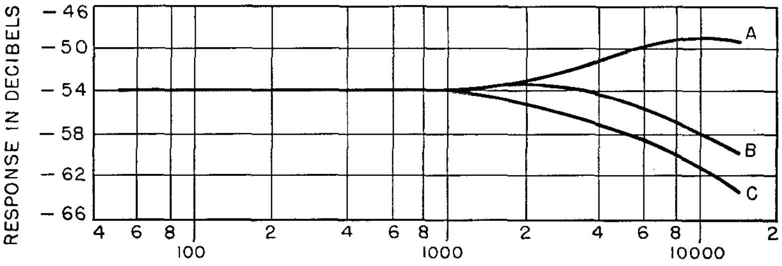

Figure 7—Characteristic of type-*B* microphone under the same conditions as in Figure 6.

phone has a slightly greater electrode spacing and a resulting lowered sensitivity to handle linearly sound pressures of 1500 to 2000 dynes per square centimeter.

As is evident from curve *C* of Figure 7, the type-*B* free-field microphone has a gradual roll-off characteristic when tested in a reciprocity coupler, showing that the first resonant mode is deliberately suppressed to provide the uniform free-field response indicated by curves *A* and *B*. A flat response characteristic in a free field is desired for high-fidelity recording, where a single capacitor microphone suitably located picks up the sounds of a large orchestra. Here, since the sounds reach the microphone at random incidence, the deviation from linearity is small and consists of very-uniform changes over a broad range of frequencies; it is undetectable audibly.

As shown in curves *A* and *B*, the response to frequencies above about 3500 cycles can be emphasized, de-emphasized, or rendered relatively flat by positioning the microphone diaphragm with respect to the incidence of the sound. This may appear as a fine point just after mention of the fact that the response is very uniform to orchestral sounds reaching the microphone in random incidence. Nevertheless, the suggestion here is concerned with close-microphone recording, where it is desired to provide some high-frequency emphasis or a slightly accentuated brilliance to an orchestra, an orchestral section, or a voice. Similarly, where de-emphasis of brilliance is desired without increasing the distance from the microphone (and thus losing presence), proper positioning may be employed. If slightly increased microphone spacing is preferred to prevent blasting, microphone orientation can be employed to recreate the timbre qualities of close placement.

Extension of the reciprocity and free-field comparison curves to 15 000 cycles shows no sharp peaks or deviations from uniformity. Carrying the investigations further by the actuator method has shown that the uniform roll-off carries on to ultrasonic frequencies on the free-field version. A secondary peak occurs at about 16 000 cycles on the type-*A* microphone and is of little concern in cavity work. Since the response–frequency characteristic is relatively uniform over the range of component frequencies of transients, the response of these microphones is excellent for highly transient sources, such as speech and musical percussive sounds of the piano, cymbals, and the like. Listening tests verify that sounds of this character are reproduced with clarity, crispness, and liveness.

The following data may be of interest. A sound pressure of 154 decibels (referred to 0.0002 dyne per square centimeter), which is the output limit of our present facilities, was applied to the microphone and no electrical failures resulted. Although the microphones were found to be linear in output voltage with respect to the applied signal, the system distortion in this test as measured on a harmonic-distortion meter for type A was about 3 percent and for type B about 4 percent when 200 volts of polarization were used. The system distortion dropped to below 1 percent for type A at 142 decibels and the similar distortion point for type B is about 134 decibels. Since the distortion values obtained were well below the American Standards Association requirement, no attempt was made to eliminate the distortion introduced by the sound source and preamplifier circuits used.[7] Therefore, our rating of the microphones based on the overload point for type A is 140 decibels and for type B, 130 decibels; values that allow a good margin for faithful reproduction. To indicate that distortion is a function of loudness of signal, when two microphones were used in a coupler, one as the sound source and the other as a microphone transmitter, the total distortion in our reciprocity apparatus was less than 1 percent at 85 decibels. This is the maximum level obtainable. Assuming that half of the distortion was associated with each microphone, the distortion at this level was considerably less than 0.5 percent.

The insulation resistance of the microphones with 250 volts of direct current applied is over 200 000 megohms for normal room conditions and has been found to remain at that figure indefinitely. Type-A microphones usually withstand direct voltages at 385 volts and type B of 270 volts. Individual microphones deviate from the usual breakdown values within -10 to $+20$ volts. The recommended polarizing voltage is 200 ± 20 volts, which provides an adequate safety factor for a simultaneously applied polarizing voltage, loud signal, and physical jolt.

As a result of adequate factors of safety, namely: sufficient breakdown voltage, high insulation resistance, tensioned diaphragm, and special attention to cleanliness in manufacture: a low value of inherent noise is obtained. Measurements to obtain the inherent noise of the microphone showed that it is below the measuring capabilities of our equipment. The noise level of the preamplifier was below 0.0001 volt and this below-scale reading was not altered with or without the microphone in the circuit. This value is in agreement with published information on such a device, which for inherent noise may be considered as a 40-micromicrofarad capacitor shunted by a resistor of 20 megohms.[8]

These microphones, when cycled through temperature changes from -50 to $+60$ degrees centigrade, were found to maintain calibration accuracy. Whereas the American Standards Association specifications allow 0.05-decibel change per centigrade degree, the microphones appeared to vary only 0.02 decibel or less. This value cannot be presented with a higher degree of certainty at present since it is difficult to extract all effects of temperature variations occurring in the associated equipment.

The construction is such that when the rear of the microphone is open to the pressure existing on the front of the diaphragm, equalization to changes in pressure results. In laboratory tests, a pressure change from 10 to 740 millimeters of mercury over a 10-second period caused no visible distortion of the diaphragm. However, this pressure change during a 5-second interval caused the diaphragm to touch the rear electrode. Although this treatment did not affect the performance of the microphone, it is recommended that such changes be extended over a period of time somewhat greater than 10 seconds. It is further recommended that the polarizing voltage be removed during extensive pressure changes to prevent possible sparking and resulting electrical deterioration of the microphone.

4. Performance Limitations

For accurate work, certain precautions are important when using capacitor microphones. Since they have the characteristics of high-impedance electrical devices and use air for the dielectric and working medium, corrections should be made when environmental influences are abnormal. Altitude changes affect the acoustic controls, and corrections for the altered response–frequency characteristics must be made when large changes are involved. When the microphones are used for calibration work requiring greatest accuracy, corrections for these variables and in some cases for the capacitance change should be applied. If not abused, the microphones are stable and have maintained calibration over many years. For all ordinary usages, such as in factory and laboratory equipments and in broadcast and recording work, the Kellogg microphones remain uniform. There is no need to store these microphones in a desiccator except possibly for long-time (5 years or more) storage of verification standards.

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[7] L. L. Beranek, footnote reference 3, page 220.

[8] L. L. Beranek, footnote reference 3, pages 221–223.

Some capacitor-microphone characteristics are useful indicators of questionable performance. Should the calibration or sensitivity shift sharply, the microphone should be suspected of being defective due to mishandling or for other reasons. A measurement of the insulation resistance (through a 1-megohm protective resistor) and of its capacitance should be made. If the insulation resistance changes are 15 percent or less, the damage is slight and the microphone is still serviceable. Each micromicrofarad variation in capacitance changes the sensitivity by about 0.1 decibel and if the change exceeds 4 micromicrofarads the sensitivity and frequency characteristic will be seriously altered. An unsatisfactory microphone is usually characterized by flashovers or noise, which make it necessary to remove the microphone from service—usually before response–frequency deterioration would be suspected.

5. *Conclusions*

The stretched-diaphragm capacitor microphone has been recognized for a long time as having a very-uniform response–frequency characteristic, a wide dynamic range, low distortion, and low noise. Present-day recording techniques and higher-quality reproducing systems demand these performance advantages. The microphones described herein meet these requirements and also provide a good degree of ruggedness and calibration stability without sacrificing desirable characteristics. Further, the bandwidth of the microphone has been extended by allowing slight deviations from linearity. Since these deviations are minor and are spread over a broad range, they are not audibly detectable. Studies of response–frequency characteristics show the particular designs to be advantageous and emphasize the necessity for the two types of microphones, each specialized for its field of application.

6. *Acknowledgment*

Indebtedness is acknowledged to Mr. C. M. Miller, who is responsible for much of the design and exacting assembly of the microphones, and to Messrs. R. V. Hagen and C. H. Royce, who participated in the more-recent technical and refinement studies. The research and development were performed under the general direction of Dr. G. H. Brodie, who along with the others aided in the preparation and editing of this paper.

Manufacture of Type-7400 Step-by-Step Telephone Selector

By J. L. DONCEEL

Bell Telephone Manufacturing Company; Antwerp, Belgium

THE NEW TYPE-7400 step-by-step selector has been developed for use in telephone- and teleprinter-type machine-switching exchanges.

The selector is a 30-point single-motion rotary switch with wipers arranged to provide 30 or 60 outlets. The brush carriage with ratchet wheel is actuated by a stepping mechanism operated by self-interruption. This step-by-step selector incorporates the following improvements over previous designs.

A. Greater stability of operation.

B. Greatly reduced maintenance.

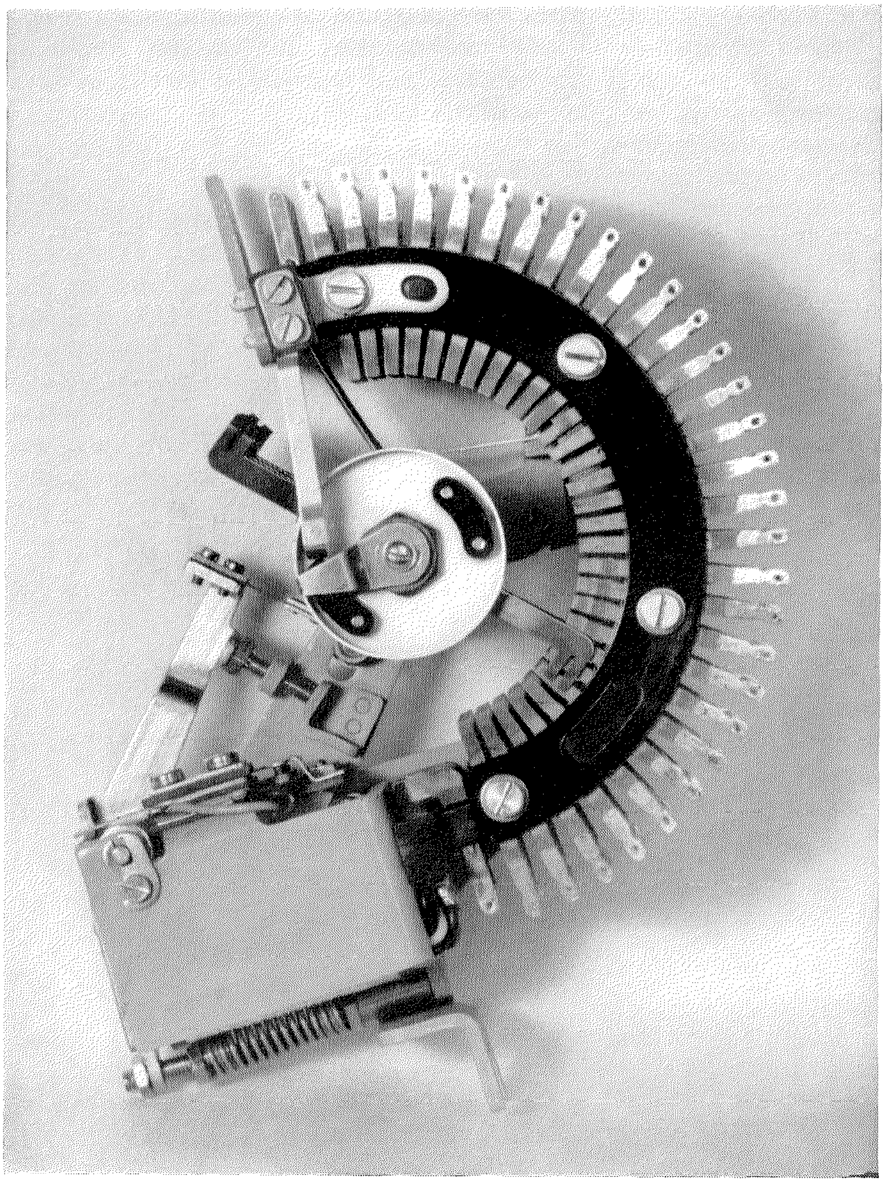
C. Interchangeability of parts.

D. Stepping mechanism and rotor is a jack-in unit that can be adjusted before mounting on the contact bank.

E. Parts are very easily accessible and in plain view, allowing a visual check of operation.

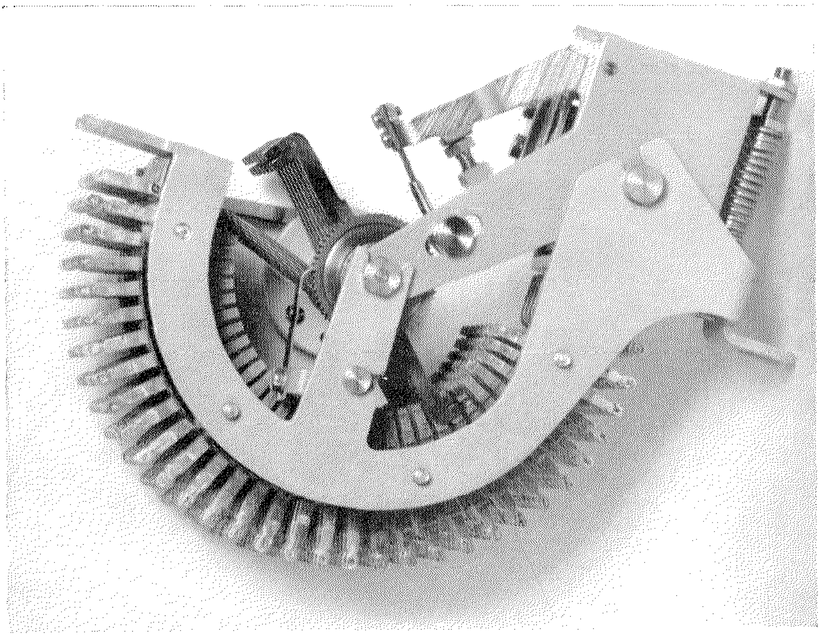
The different piece parts are manufactured within very accurate limits to ensure interchangeability of components.

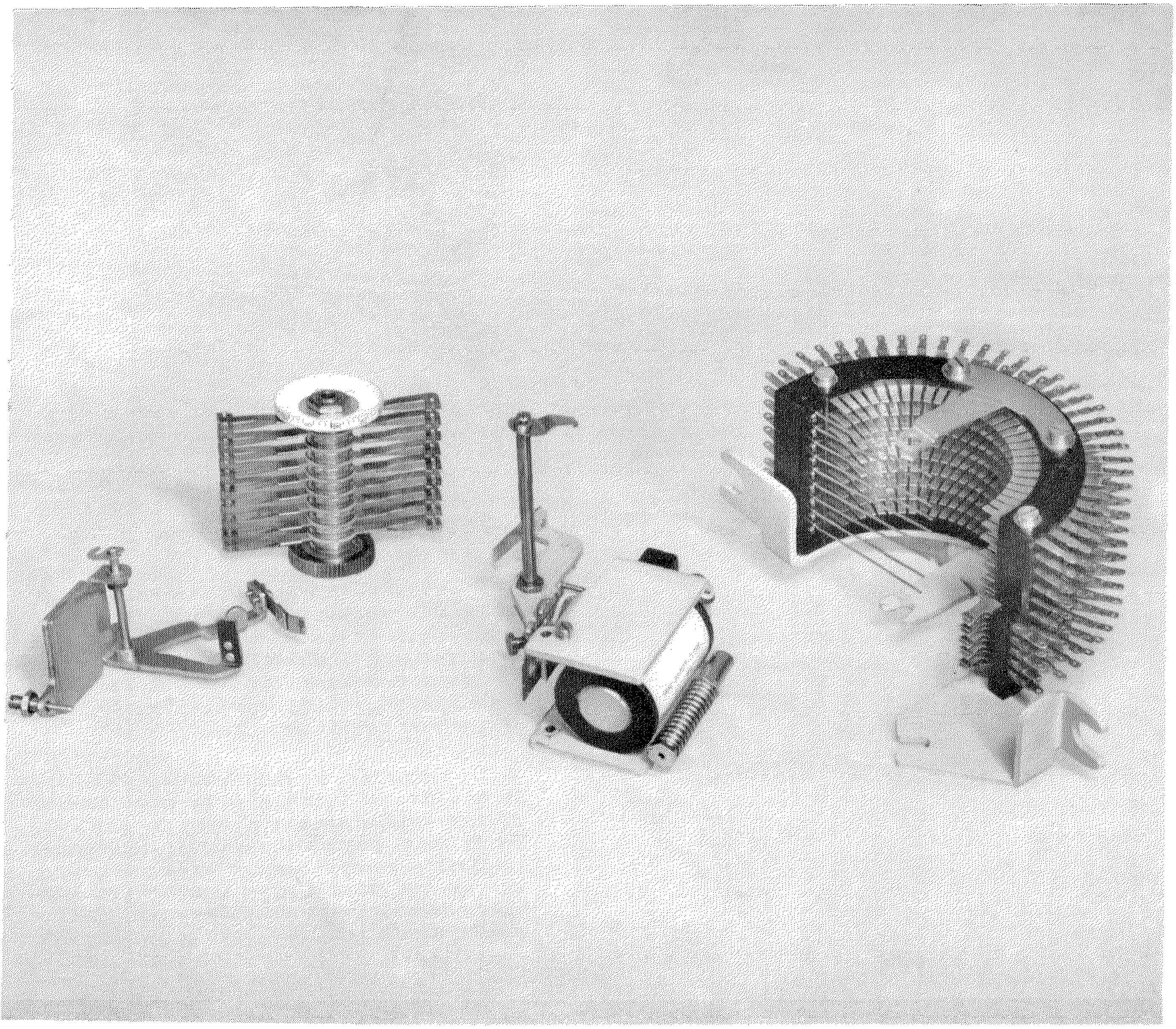
The maximum capacity of the selector is 10×30 points or 6×60 points. The switch operates at normal potential at a speed of 60 steps per second under self-interruption.

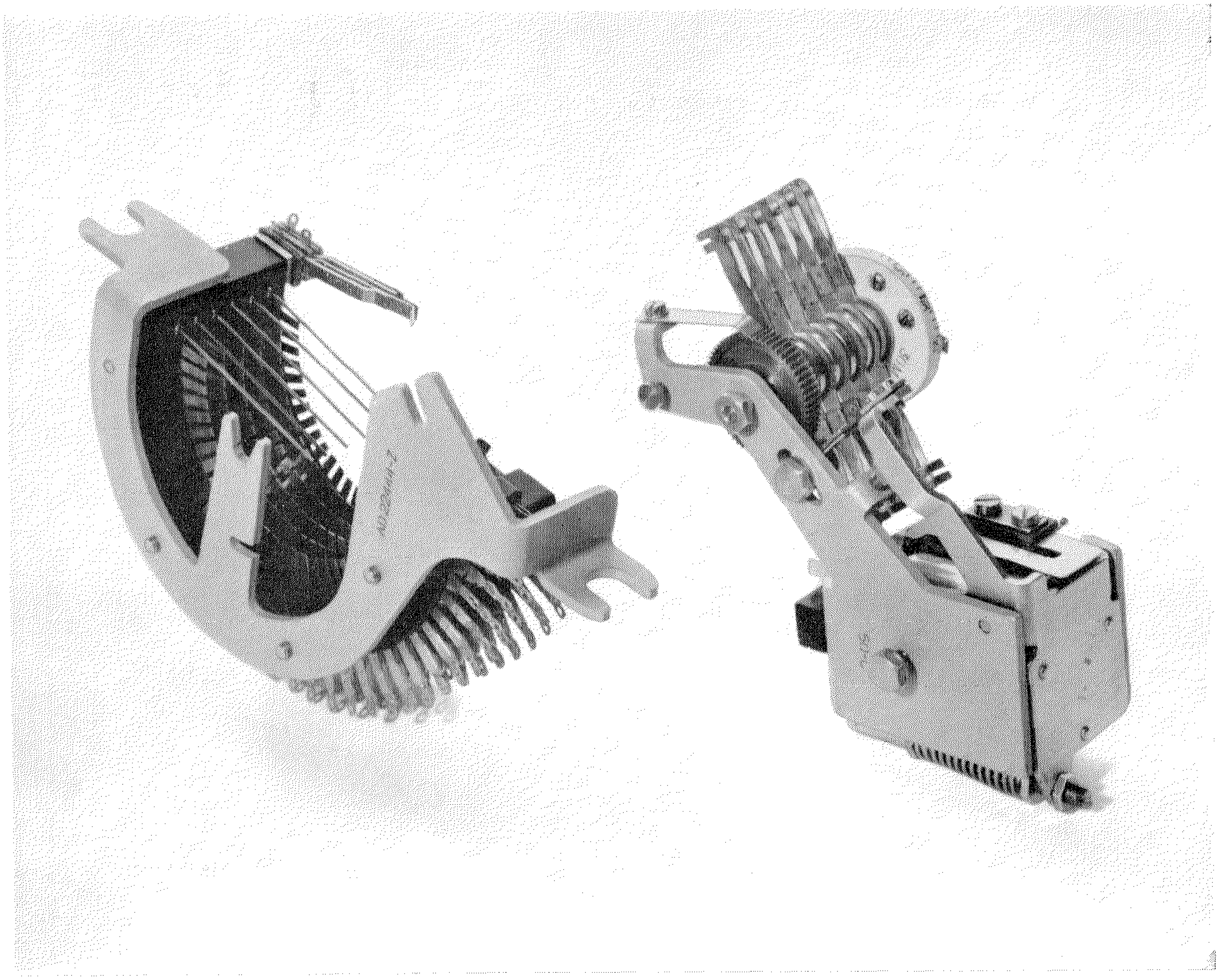


1. The assembled selector.

2. Another view of the selector from the base-plate side is given at right. Below, the major assemblies of the selector are, from left to right, the armature with driving pawl and adjusting screw for the armature spring; the rotor; the yoke with retaining pawl, armature spring, power coil, and rotor axis; and the contact bank.



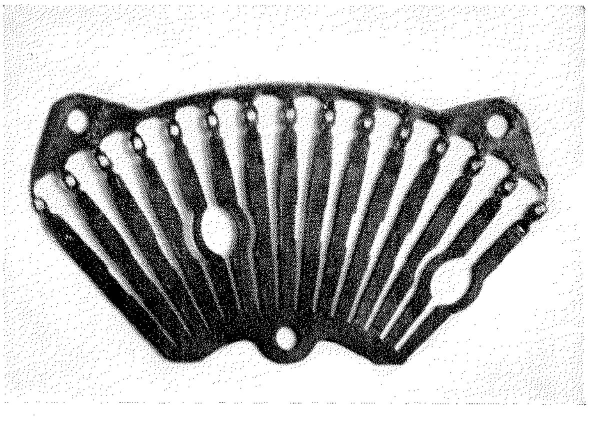




3. The selector can be disassembled into the two main parts shown here; at the left is the contact bank with the off-normal contact and, at the right, the stepping mechanism with the rotor.

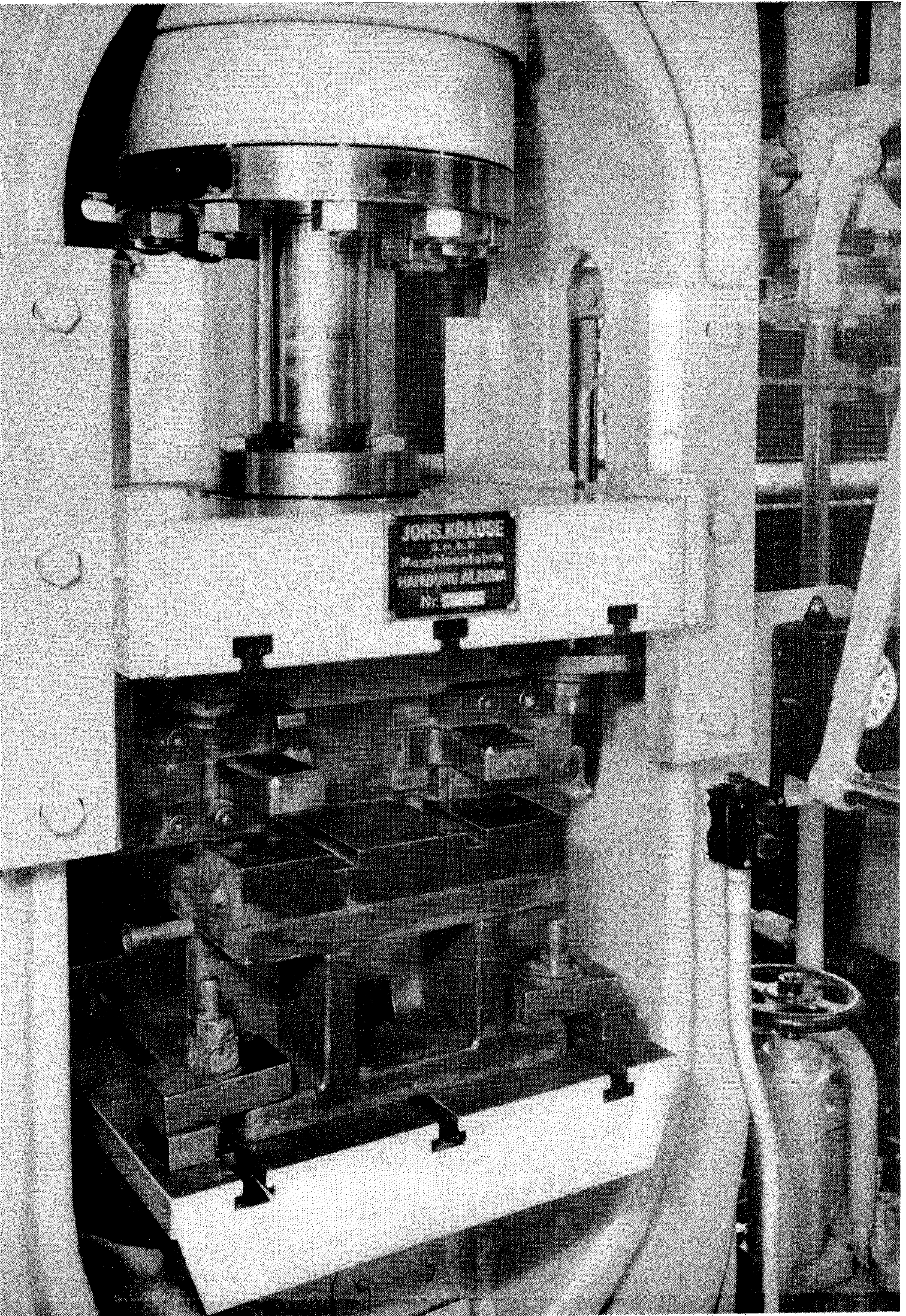
Each of these two parts forms a complete unit that is preadjusted by means of suitable gages, after which they may be assembled easily and rapidly without further adjustment. The stepping mechanism is attached to the bank by one screw.

The jack-in connection between the winding of the power magnet and the interrupter on the bank greatly facilitates maintenance work.



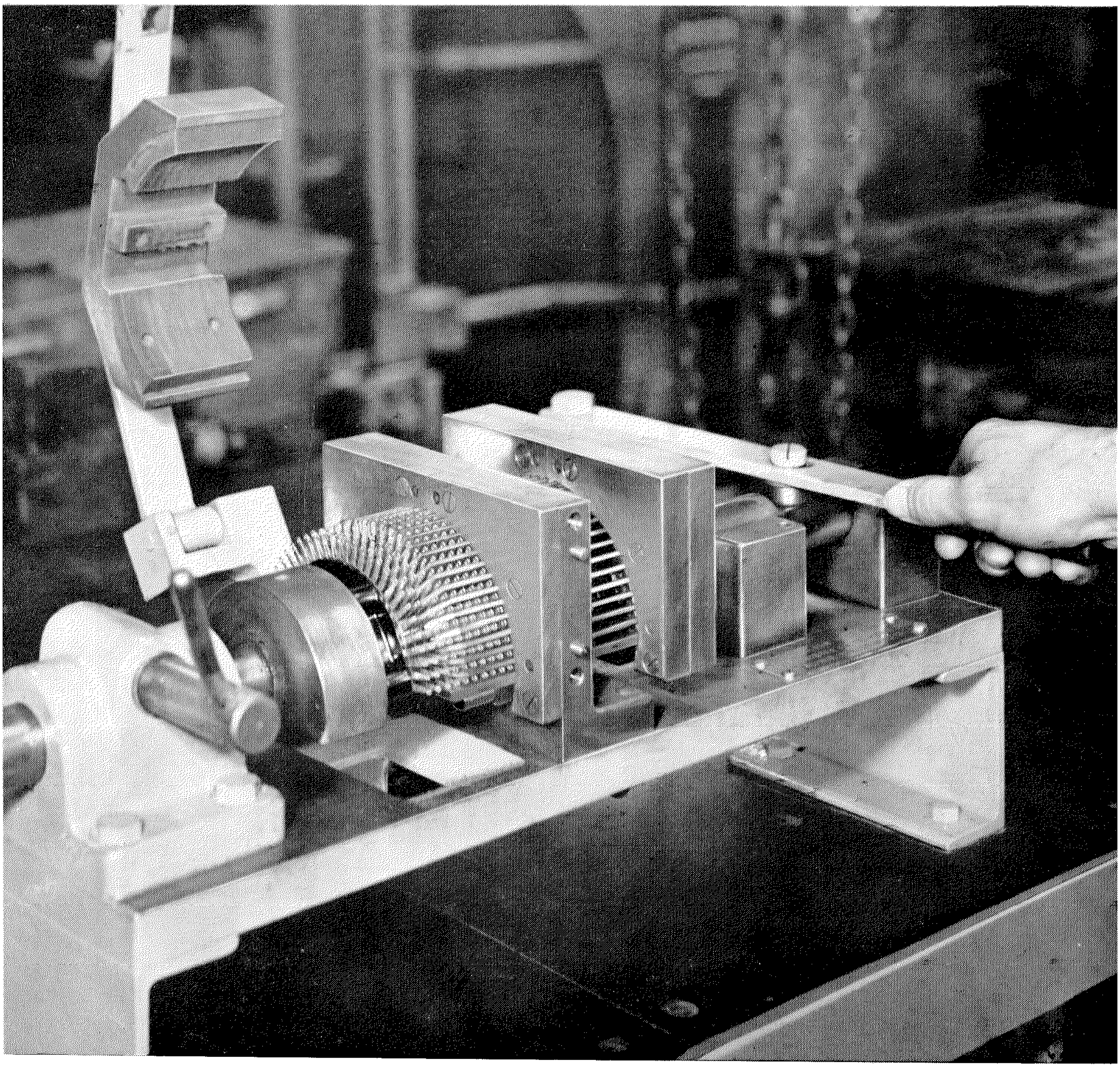
4. The arc terminals of the contact bank are of phosphor bronze punched out in a form of a sector, an example of which is shown in the photograph at the left. Each sector incorporates 15 terminals. These are stacked in various combinations to provide the desired number of levels and contacts in the molded bank.

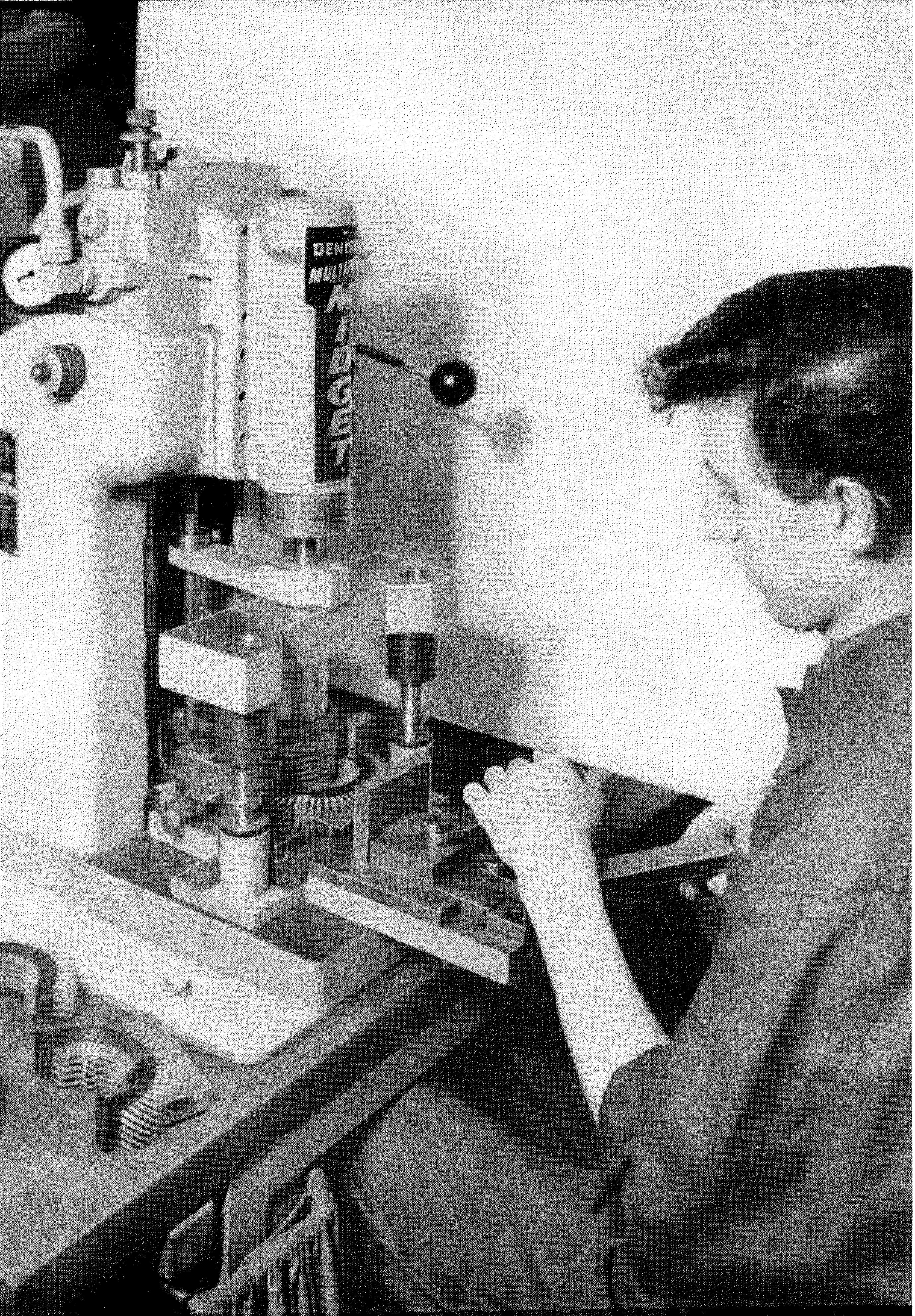
5. The contact bank is a single injection-molded bakelite unit. The mold is designed so that contact banks of different capacities may be made using the same mold. For the insulating material between the terminals, preference has been given to a thermosetting compound, as this gives the best guarantee against temperature distortion and may be molded within sufficiently accurate limits. The molding press in operation is shown on the facing page.


JOHS. KRAUSE
G. m. b. H.
Maschinenfabrik
HAMBURG-ALTONA
Nr.

6. Below, after the molding operation, excess bakelite is removed by a suitable tool. All bakelite particles between the terminals are removed in a single operation.

7. The next process is cutting out the metal sectors t hold the terminals together during molding. This is d in a midget multipress as shown on the facing page.

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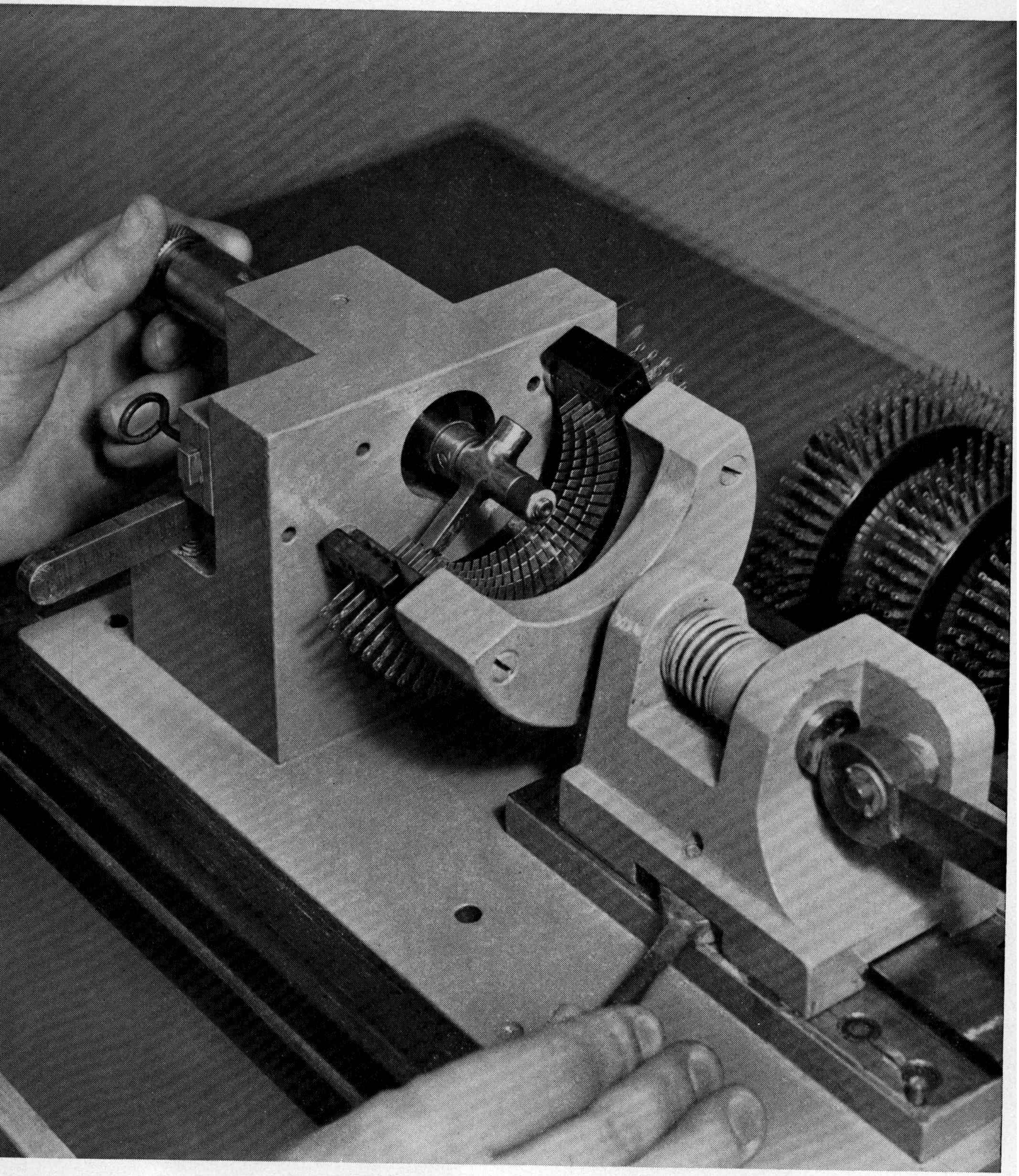
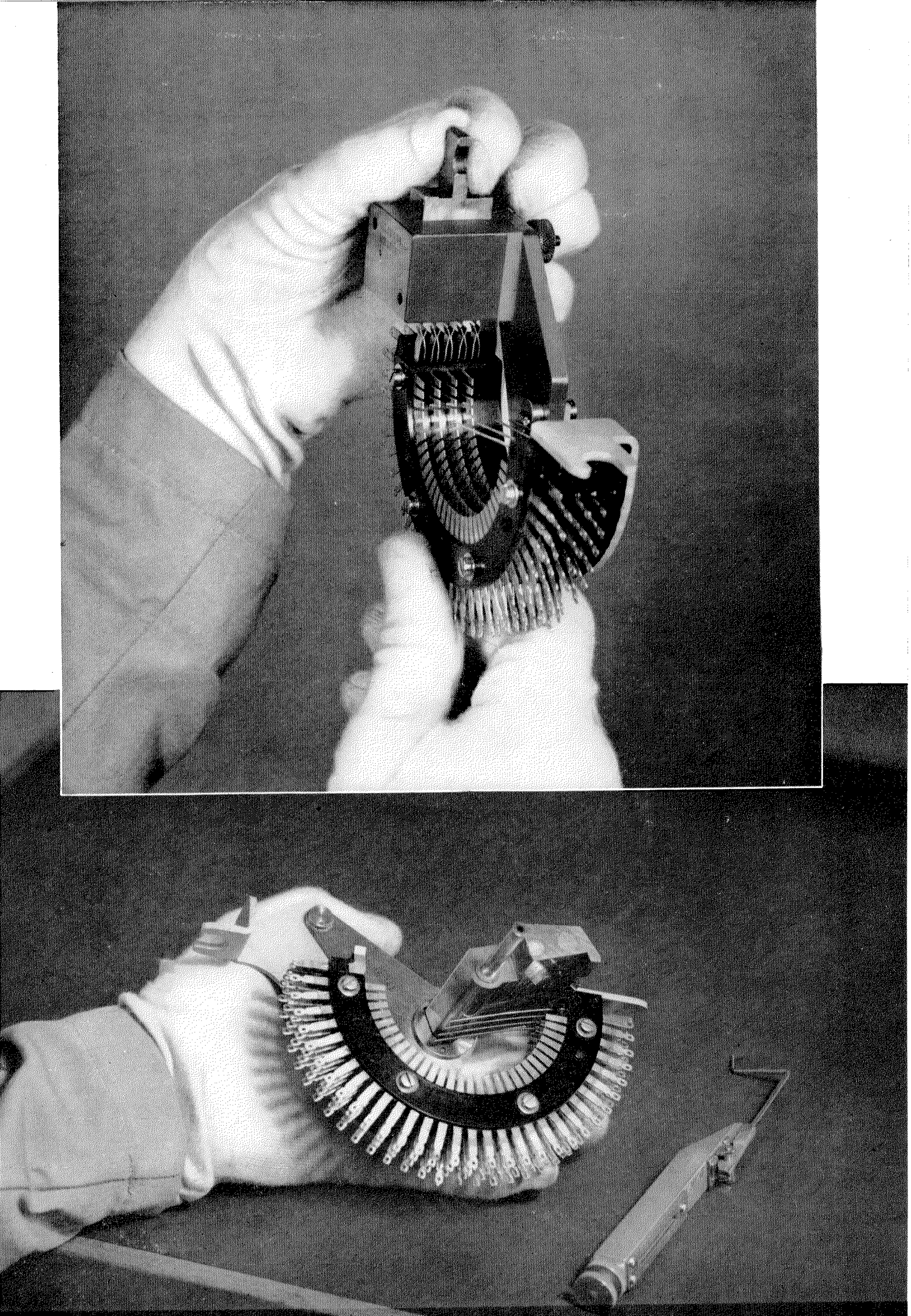
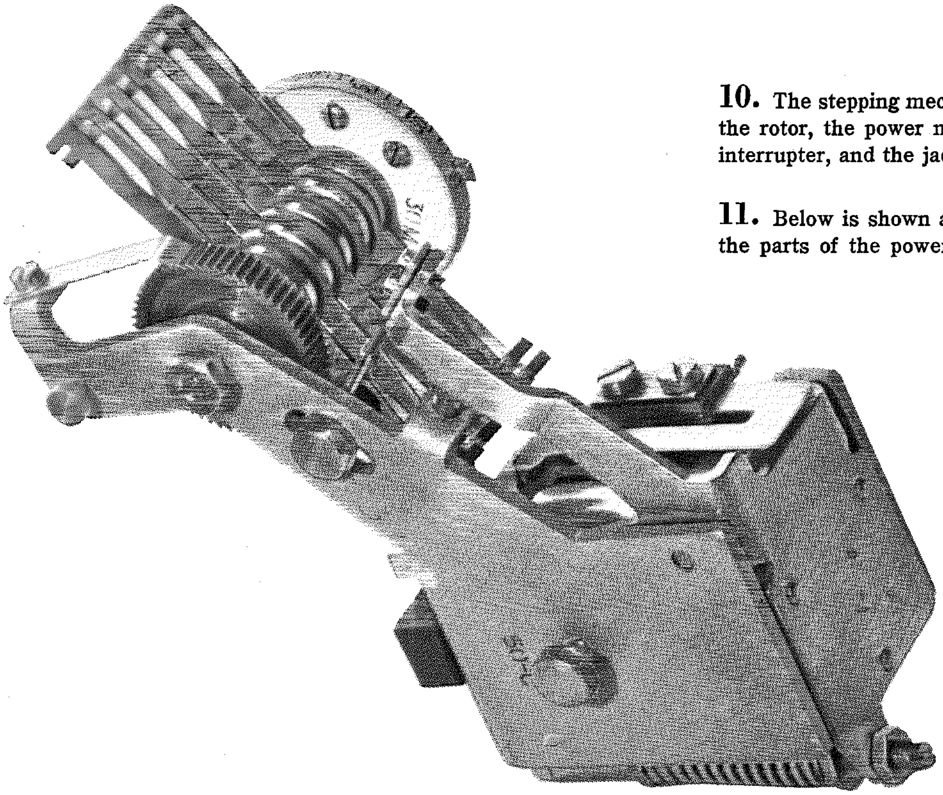


8. The contact banks are then checked for correct position and alignment of each level of terminals. This is performed in a single operation by the special gage shown in the photograph above.

9. The frame holding the contact bank is a simple flat punched-steel plate with mounting lugs bent over at 90 degrees. The frame is simply manufactured and is punched within very accurate limits with no supplementary treatment afterward. It is dimensionally stable; deformations due to ageing are eliminated.

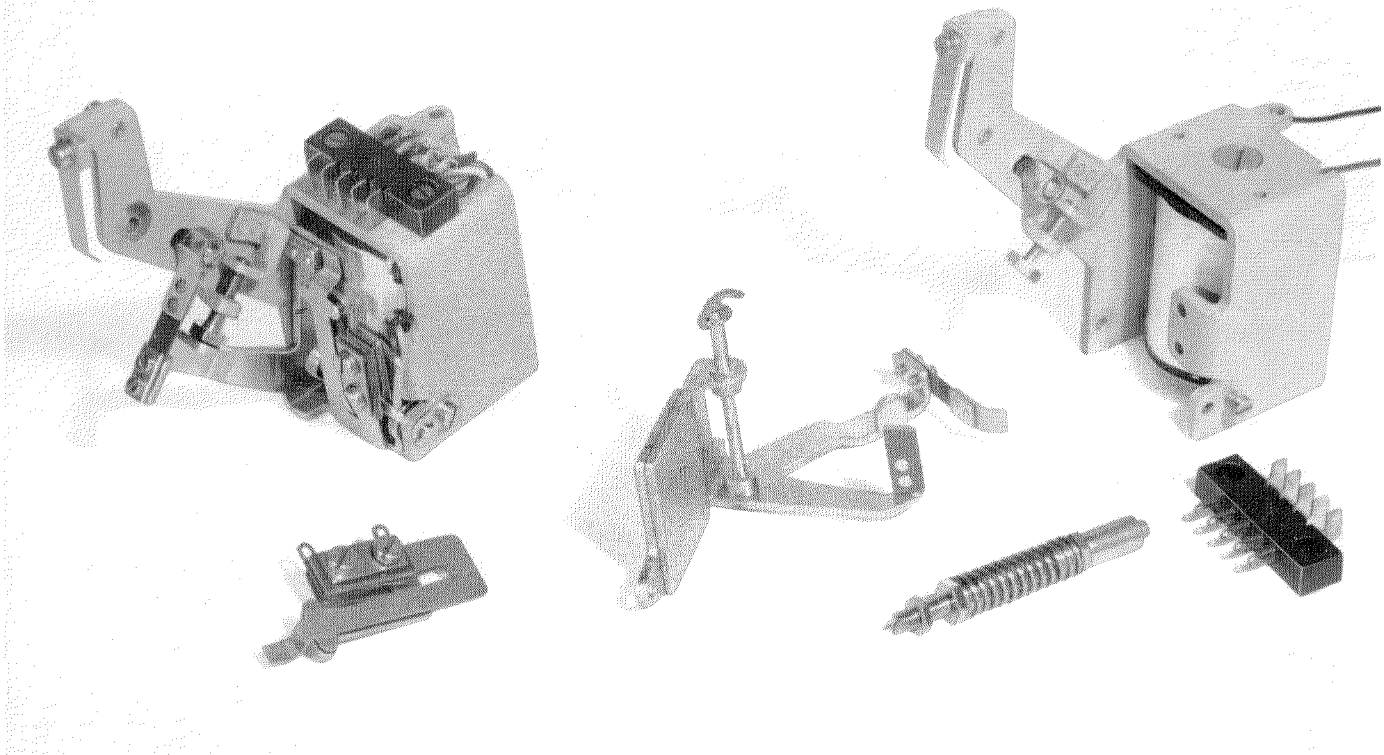
The contact bank is now assembled into the frame and the collector springs and jack-in springs are mounted on the contact bank. The springs are located by slots in the molding into which they are inserted and fixed by clipping. This permits the springs to be easily removed from the arc for possible replacement. The facing page shows how the collector and jack-in springs are checked and adjusted.





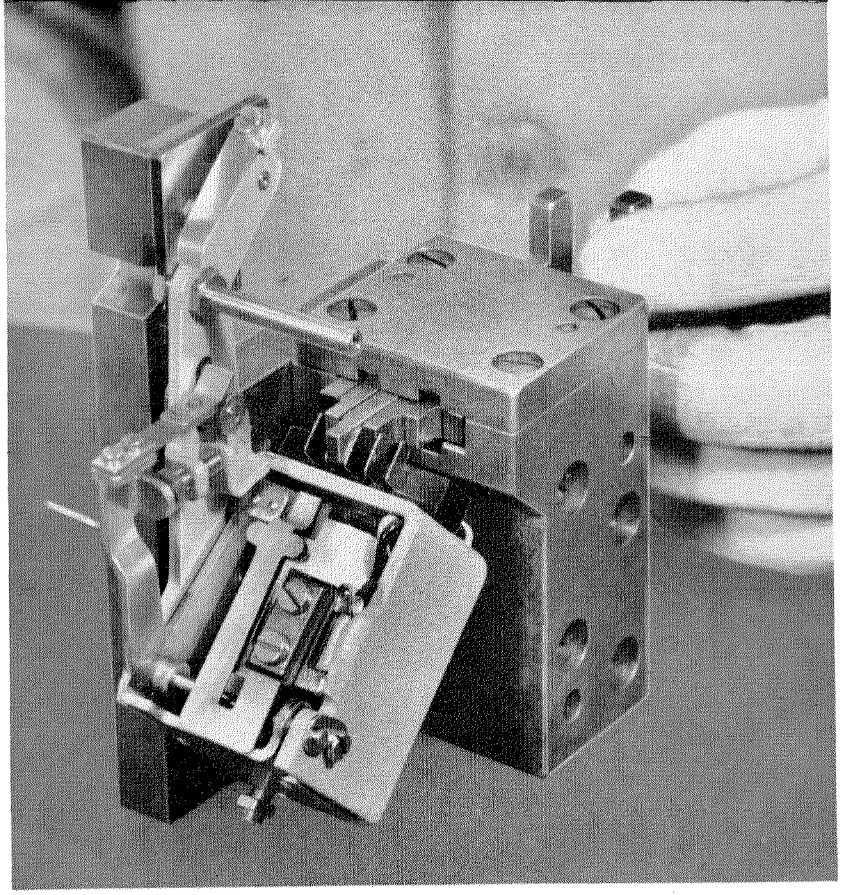
10. The stepping mechanism shown at the left comprises the rotor, the power magnet with driving pawl, the self-interrupter, and the jack-in contacts.

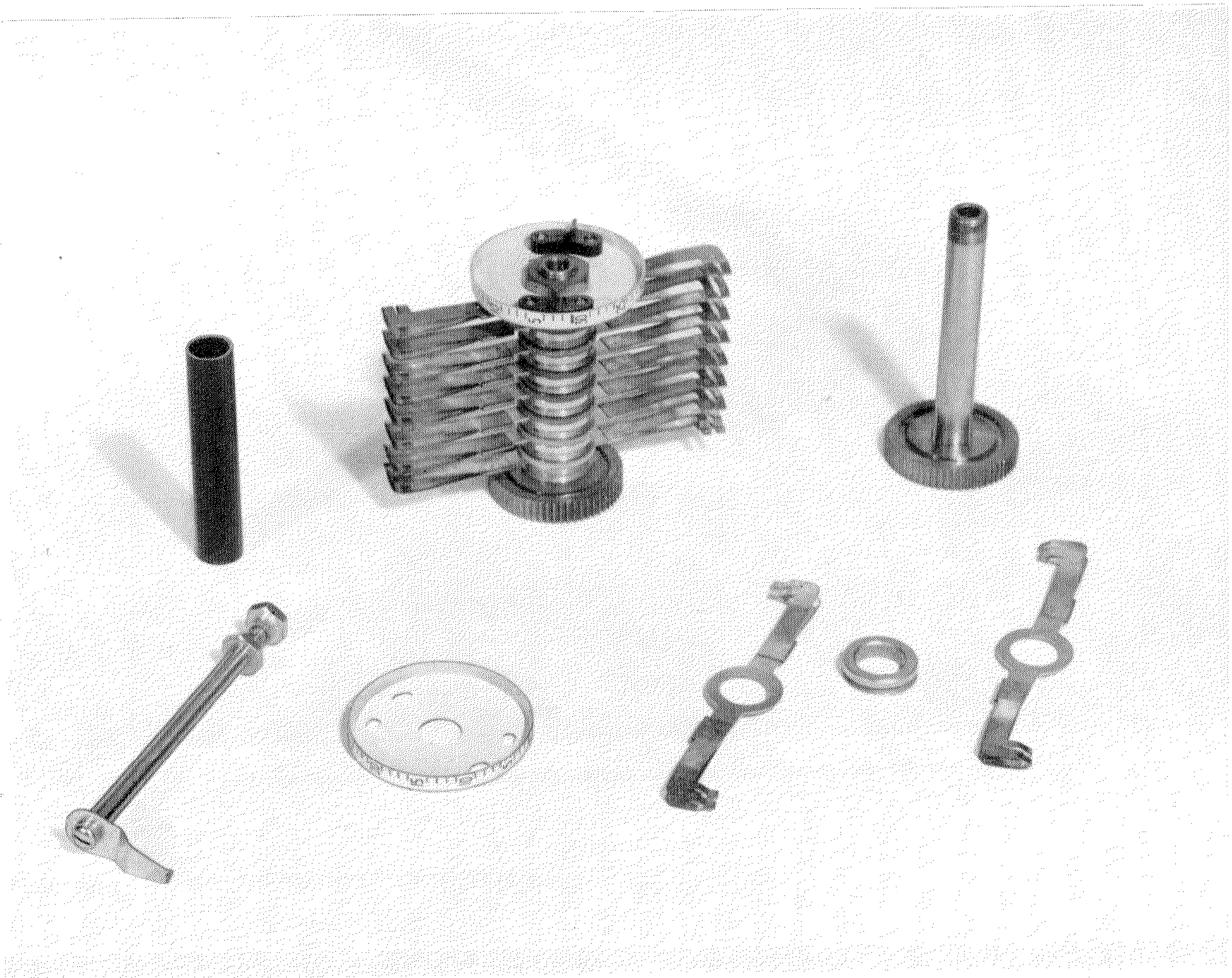
11. Below is shown an assembled view (upper left) and the parts of the power magnet with self-interrupter and jack-in contacts. The armature may easily be removed without dismounting any other part of the switch. The main spring has therefore been made of the push type and is provided with a self-locking screw for accurate adjustment of armature tension.

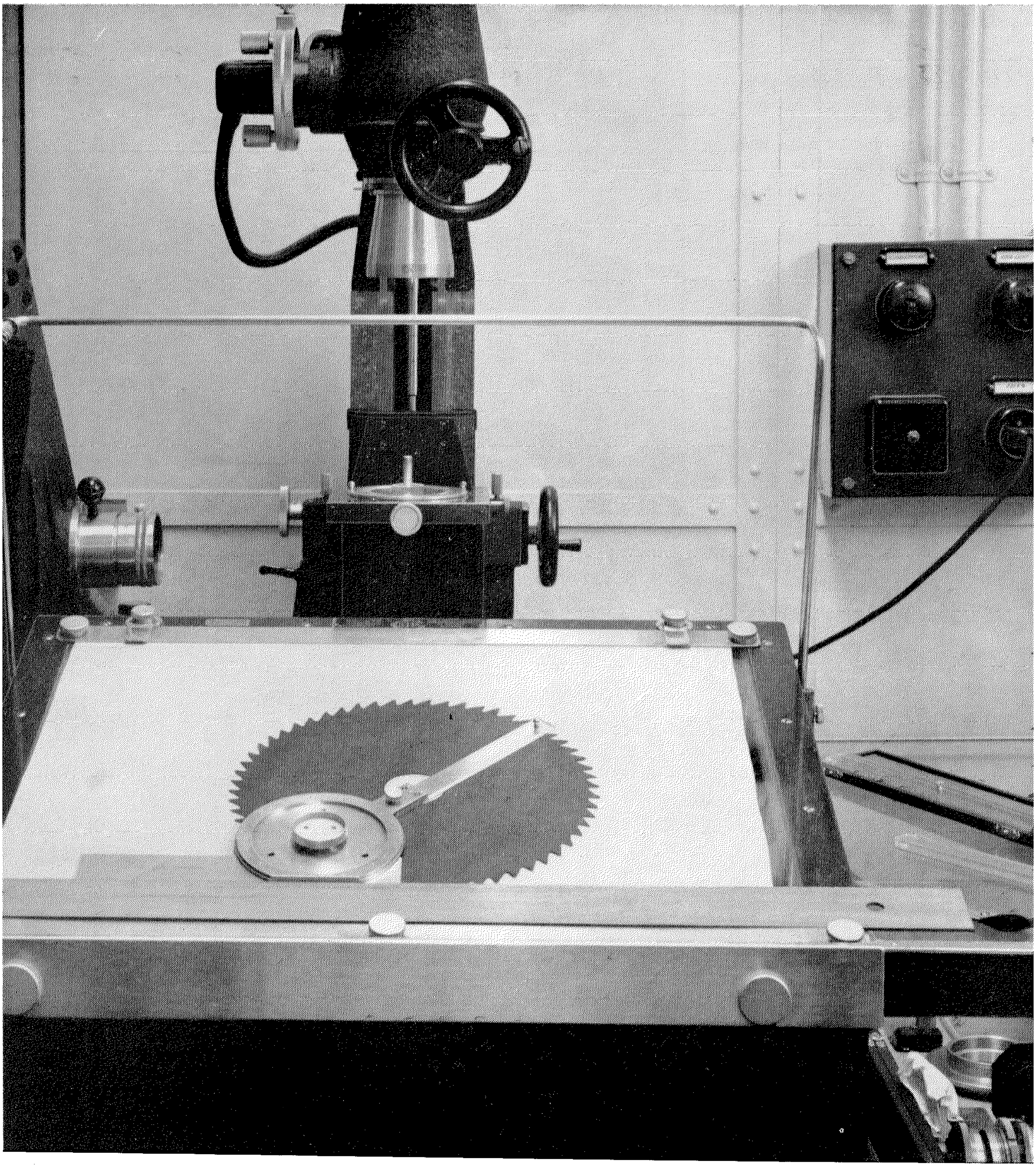


12. At the right is shown the gage for checking the correct position of the jack-in contacts.

13. Below, the assembled rotor and its parts, consisting of a ratchet wheel with hub on which the contact brushes, collector rings, and number wheel are mounted.

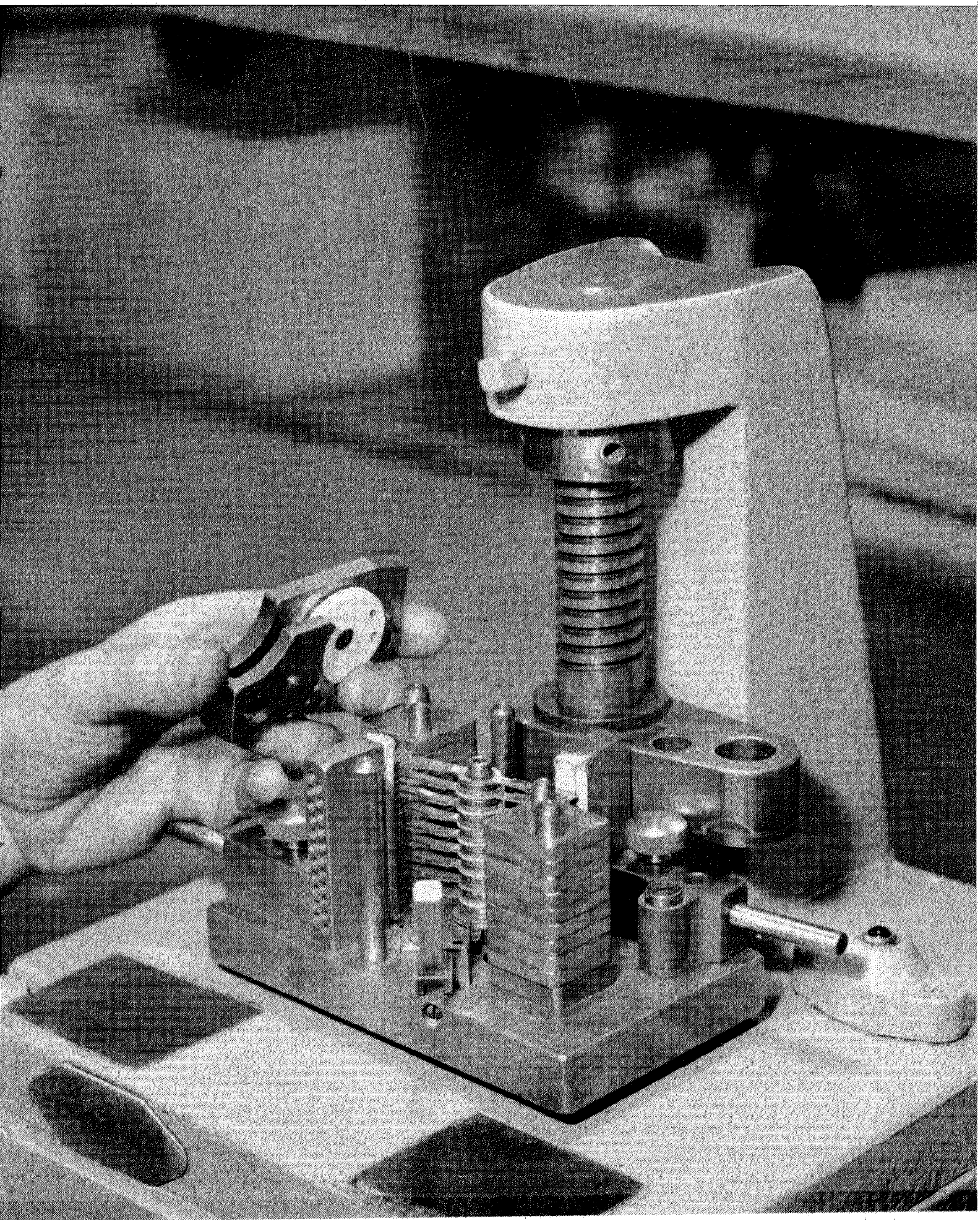


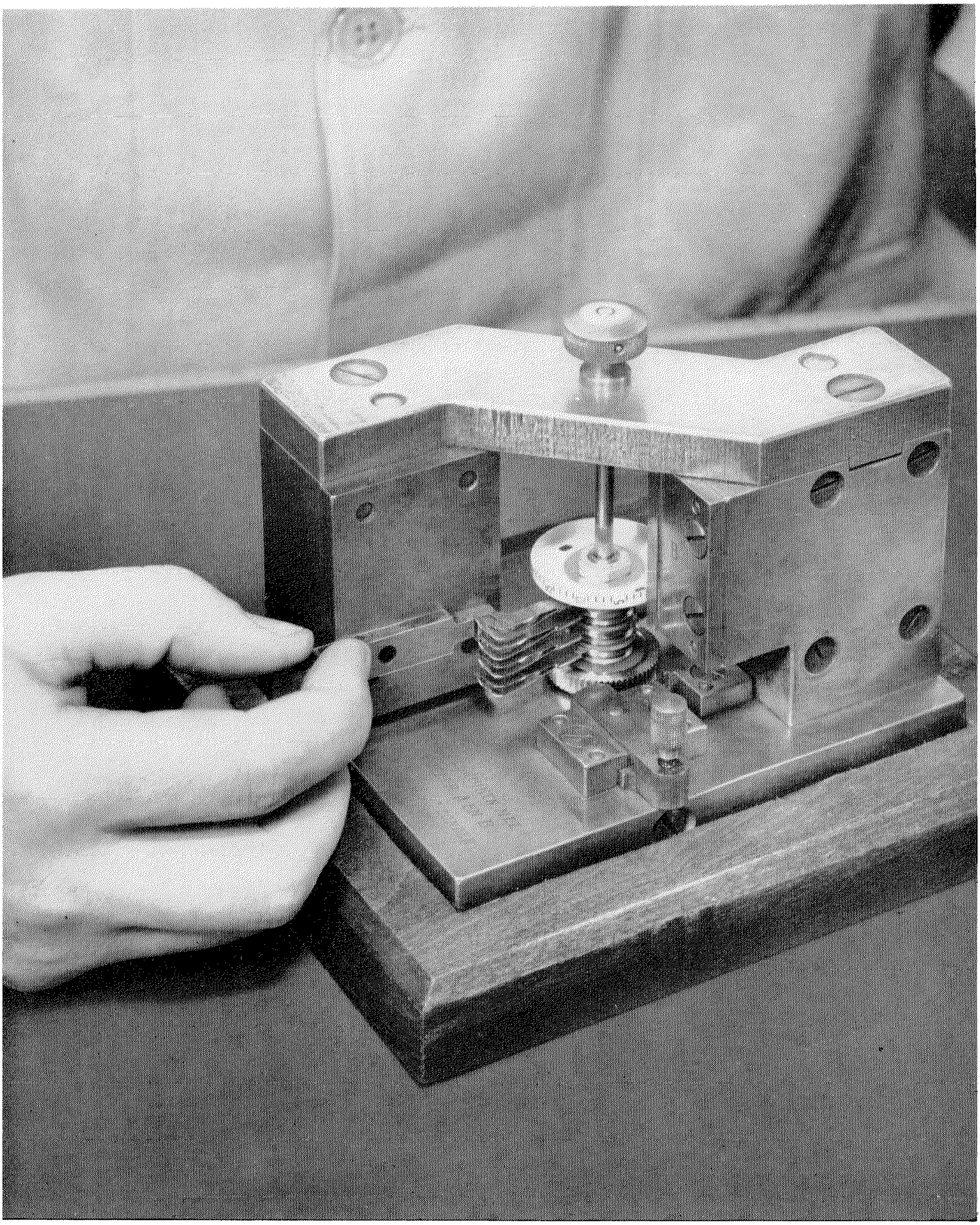




14. The teeth of the ratchet are checked on an optical comparator to ensure correct operation.

15. The rotor is assembled by means of the fixture shown here and is afterwards squeezed under a weight of 200 kilograms (4400 pounds) for firm fixation of the parts.



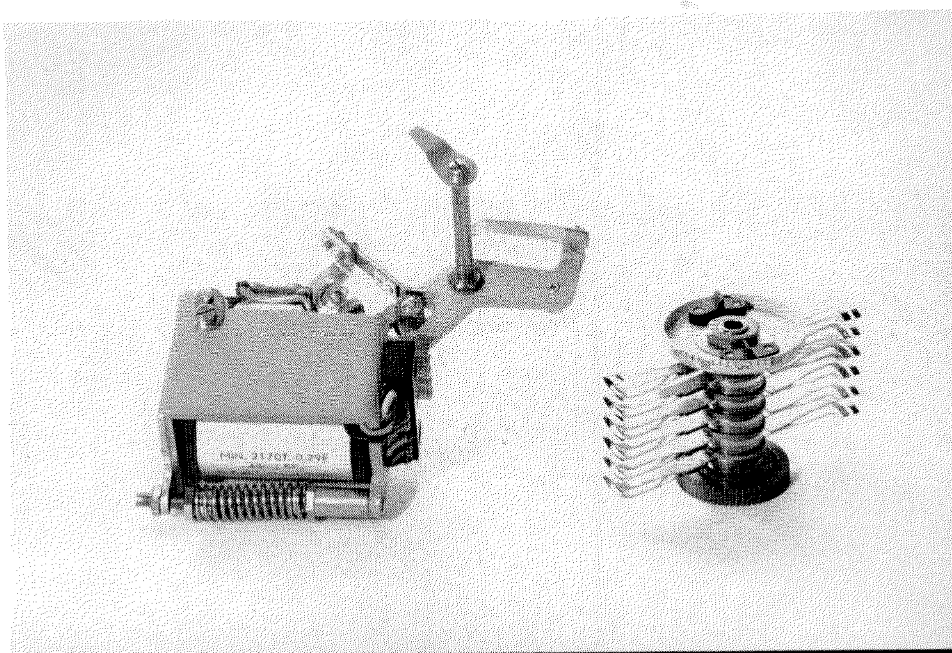


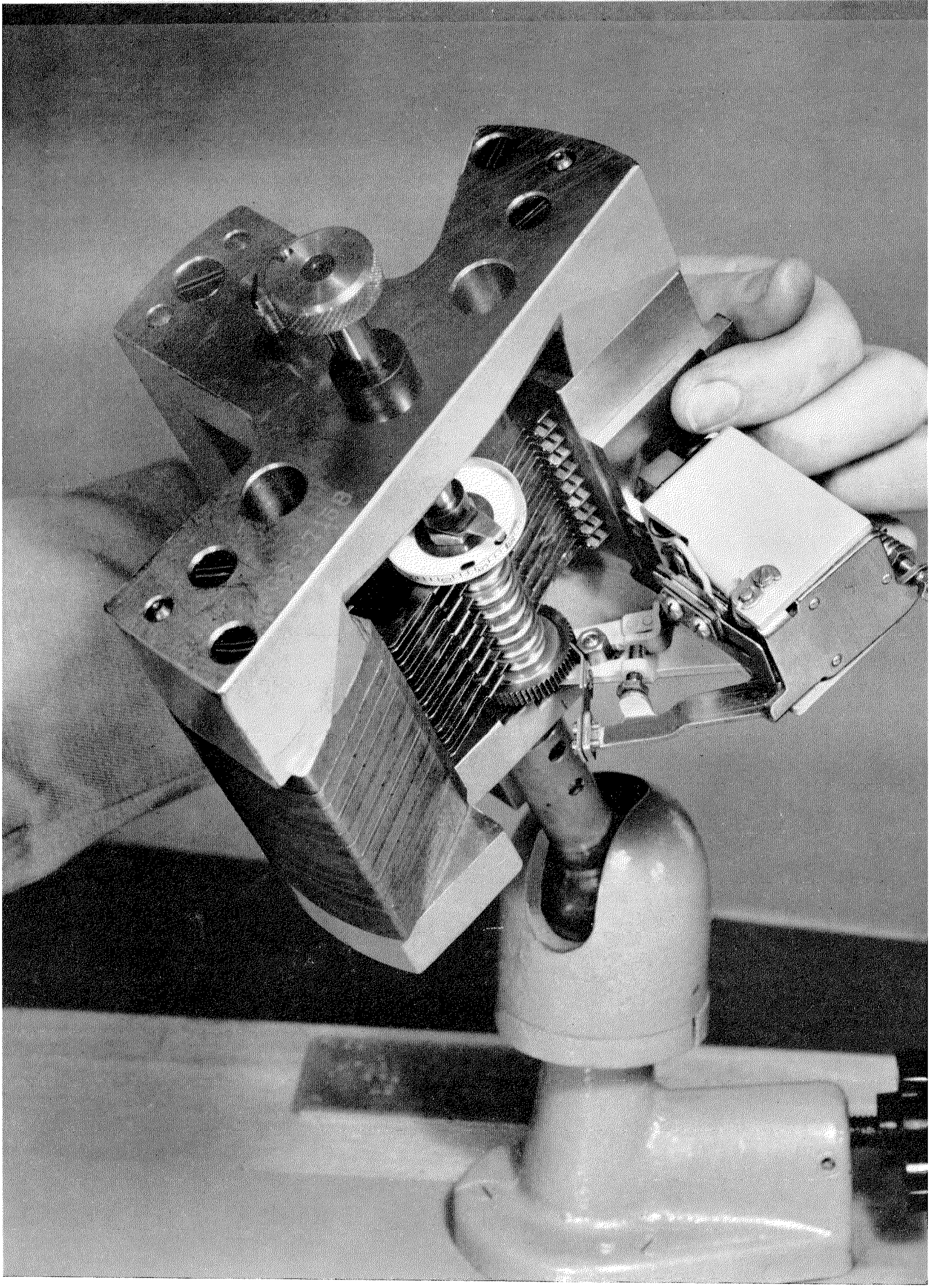
16. The contact brushes are of a special construction and are made in such a way that they may be adjusted to run properly over the terminals without affecting the contact pressure of the springs. Both springs of the same brush are therefore welded to a brass brush member that is fixed in the pile-up of the rotor and may be adjusted to obtain the correct brush alignment.

After assembly of the rotor, the correct alignment of the spring tips of the different brushes of a set is checked on a gage as shown here. The position of the number wheel is checked at the same time.

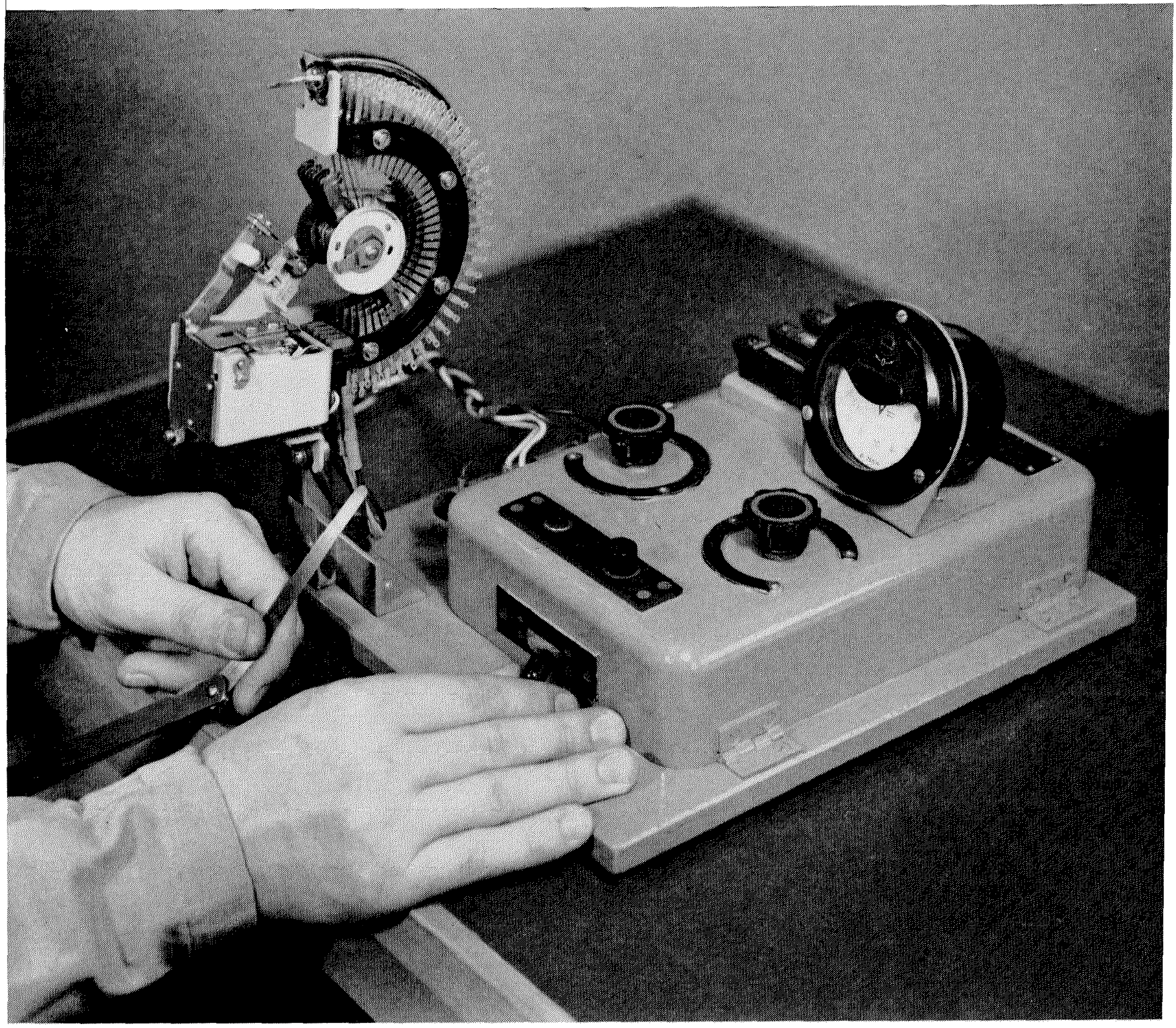
17. Above, the rotor is adjusted for the contact-brush position and for contact pressure on a gage. After this, the rotor may be mounted on any switch without further adjustment.

18. The photograph at the right shows the power magnet and assembled rotor.





19. On facing page, the magnet-and-rotor assembly is checked by a gage that represents a theoretical contact bank. This verification is of particular importance to obtain the interchangeability of the stepping mechanism to any contact bank of regular production.



20. When the contact bank is assembled with the stepping mechanism, an electrical test is made. It consists of a current test for armature operation and for correct operation under self-interruption. *Above is shown the test box and fixture for the step-by-step selector.*

Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors *

By J. W. HOWES

Standard Telephones and Cables Pty. Limited; Sydney, Australia

BASIC properties of thermistors, the terms used to specify their characteristics, and some of their applications to electronic and telecommunication problems are described in this tutorial paper.

• • •

1. *Properties of Thermistors*

Thermally sensitive resistive materials have been known to scientists for over a hundred years. As early as 1834, Michael Faraday reported measurements on the high negative temperature coefficient of resistance of silver sulphide and later experiments by English and German scientists, using magnesium–titanium oxides and the oxides of uranium, have been recorded. The temperature coefficients obtained in this early work were between −1.5 per cent and −2.5 per cent per degree centigrade at 20 degrees centigrade, but the results obtained varied from sample to sample and moreover the characteristics proved unstable in life, particularly after exposure to moderate temperatures or to the passage of electric current through the sample.

Modern development may be considered to date from about 1930, when the Bell Telephone Laboratories sought a device to compensate for the variation in resistance of telephone transmission lines with temperature and concentrated on the study of a few carefully selected materials. After several years of research and development, all of the previous disadvantages had been overcome, and semiconductor materials had been produced having extremely stable characteristics and long life. Laboratories in other countries were also active and finally a new circuit device, in the form of an extremely stable thermally sensitive resistor was made available to industry. Various trade names were adopted by different manufacturers, one of the best known being *Thermistor*, which is a contraction of the words *thermally sensitive resistor*, and for convenience this name will be used in the present paper.

1.1 COMPOSITION AND MANUFACTURE

Thermistors are complex metallic-oxide compounds usually made from the oxides of manganese, nickel, copper, and cobalt, although some manufacturers use mixed crystals such as $MgAl_2O_4$ and $ZnTiO_4$. Uranium oxide is also used, but thermistors made from this material exhibit polarisation effects and can therefore only be used on alternating current. In all cases, the desired resistivity and temperature coefficient are controlled by suitably proportioning the mixture. During manufacture, the thermistor material is fired at a high temperature and is not affected by subsequent heating up to 500 degrees centigrade, although due to limitations imposed by soldered connections, glass-bulb enclosures, et cetera, the maximum usable temperature of the finished product is about 300 degrees centigrade for those sealed into a vacuum or gas-filled enclosure and about 120 degrees centigrade for those exposed to the air.

The resistance material may be formed into small bead-type elements, tubes, rods, blocks, disks, or flakes and the heating of the element may be achieved either by passing current through the resistive material itself (directly heated type), by an independent heater insulated electrically from the element (indirectly heated type), or by placing the element in a heated enclosure such as an oven.

1.2 TEMPERATURE COEFFICIENT[1]

The resistance R decreases with increasing absolute temperature T (see Figure 1) according

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* Reprinted from *Proceedings of the Institution of Radio Engineers, Australia*, volume 13, pages 123–131; May, 1952.

[1] G. Bosson, F. Gutmann, and L. M. Simmons, "Relationship Between Resistance and Temperature of Thermistors," *Journal of Applied Physics*, volume 21, pages 1267–1268; December, 1950.

to the approximate expression

$$R = a \exp[b/T],$$

where

R = resistance at any temperature T in degrees Kelvin

a, b = constants depending on composition of thermistor material.

The temperature coefficient is $-b/T^2$ and thus decreases with increasing temperature; its value is usually of the order of -3 to -4 per cent per degree centigrade at 20 degrees centigrade compared with a value of about $+0.35$ per cent per degree centigrade for platinum.

Though the above simple law is adequate for most purposes, Bosson, Gutmann, and Simmons[1] have proposed an empirical law that gives a more accurate estimate of the resistance at any particular temperature provided that the characteristics of the material used are known.

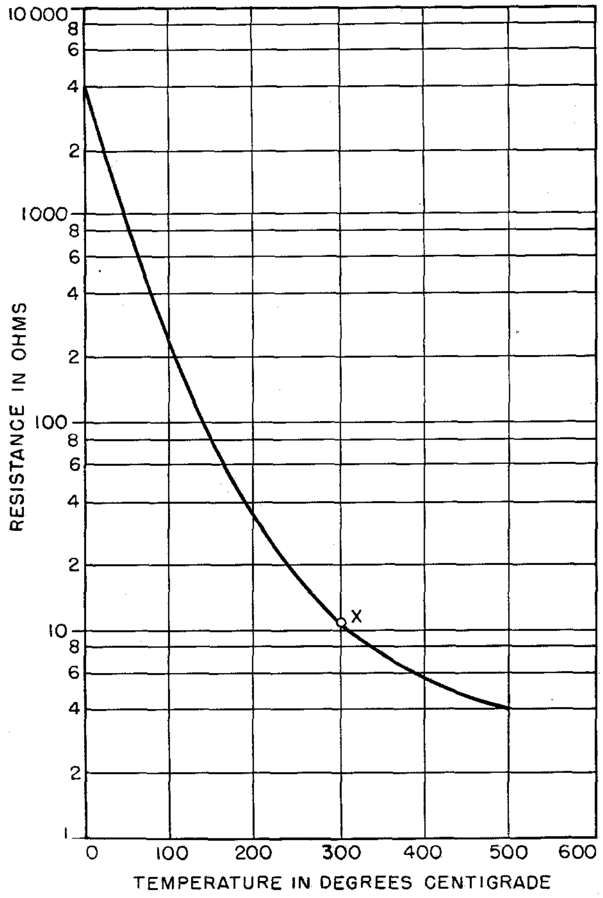

Figure 1—Characteristics of type-*F2311/300* thermistor developed for thermometry. R_c = 2000 ohms at 20 degrees centigrade. Point X marks the maximum operating temperature.

1.3 Characteristics

1.3.1 *Resistance Value*

Thermistors are available with almost any desired value of cold resistance R_c, with certain standardised values between the limits of 500 ohms and 500 000 ohms, and they may be directly or indirectly heated.

1.3.2 *Heater-Coil Resistance*

The heater-coil resistance of the indirectly heated type is generally 100 ohms ± 5 per cent and the heater efficiency is defined as the ratio between the power dissipated in the bead and the power dissipated in the heater to reduce the resistance to 2 per cent of the cold value. The heater efficiency of typical indirectly heated thermistors is approximately 90 per cent.

1.3.3 *Half-Temperature*

There are several ways of specifying the manner in which the resistance of a thermistor element changes with heating power. One way is to specify the percentage change of resistance for a change of one degree centigrade and another is by a statement of *half-temperature* or a figure derived from this. By half-temperature is meant the temperature change in degrees centigrade necessary to change the value of the resistance to a value $R/2$ or to a value $2R$. For a complete identification of the value, it is necessary to state the approximate temperature at which this half-temperature is measured. The thermistor whose characteristic is shown in Figure 1 has a half-temperature of 18 degrees centigrade at 20 degrees centigrade, whereas for the same thermistor at 200 degrees centigrade, the half-temperature is 45 degrees centigrade. At the higher temperature, when the resistance is but a fraction of its original cold resistance, a change of 45 degrees centigrade is required to produce a further halving of the resistance value, and this factor sets one limit to the useful range of operation of the thermistor.

1.3.4 Power Sensitivity

The power sensitivity is defined as the rate of change of temperature with applied power input and is expressed in degrees centigrade per milliwatt at 20 degrees centigrade. This characteristic depends on a number of factors, including the mass of resistance material and the method of mounting the resistance element. The relation between power sensitivity and power input is approximately linear over the first 50-degree-centigrade rise from room temperature, and this fact is useful in determining the effect of ambient-temperature variations on the other characteristics of a thermistor. Table 1 shows the power sensitivity of thermistors made by one manufacturer. It will be noted that the block type (*K* and *KB*) are relatively insensitive to power input and this renders them particularly suitable for use as ambient-temperature compensators. When power is dissipated by the resistance element of a thermistor it becomes hotter and thus the resistance decreases with increase of power input; therefore, when the ambient temperature falls, a greater input is required to reduce the resistance of the thermistor to any particular value. For example, it can be seen from the table of power sensitivity that a power input of 1 milliwatt will increase the temperature of a type-*A* thermistor by 3.5 degrees centigrade. Hence to maintain the thermistor at a constar resistance, 1 milliwatt more must be dissipate for each 3.5-degree-centigrade fall in ambien temperature. If trouble-free service is desirec the power input must not be allowed to excee the maximum for continuous use; this maximur is specified by the maker. For periods up to minute, it is usually quite safe to exceed thi rating by 50 per cent. In the case of indirectl heated thermistors, the total power dissipatio in the bead and heater must not exceed th maximum indicated. It should be noted that i the temperature is fixed, the thermistor obey Ohm's law; that is, the relation between voltag and current is linear.

TABLE 1

CHARACTERISTICS OF VARIOUS TYPES OF THERMISTORS

Type	Description	Power Sensitivity Degrees Centigrade per Milliwatt	Dissipation Constant Milliwatts per Degree Centigrade	Maximum Voltage Factor K
A	Bead formed on two 0.020-inch (0.5-millimetre) platinum wires and sealed in an evacuated bulb	3.5	0.28	20
B	Indirectly heated bead sealed in an evacuated envelope. Heater resistance, 100 ohms. Maximum voltage between heater and thermistor, 25 volts	5.0	0.2	22
C	Rod with pigtail ends	0.066	15.1	2.3
F	Small bead fused into tip of evacuated glass bulb	1.6 (Air) 0.4 (Water)	0.62 (Air) 2.5 (Water)	—
K	Cylinder, 0.625 inch (16 millimetres) in diameter by 0.125 inch (3 millimetres) in length	0.065	15.4	2.5
KB	Block, mounted on metallic plate 1.25 by 1 inch (6.4 by 2.54 centimetres) with four mounting holes	0.04–0.01, According to Mounting	25–100 According to Mounting	1.2–2.2, According to Mounting
L	Indirectly heated bead in a gas-filled envelope. Heater resistance, 100 ohms	2.5	0.4	16
M	Bead, sealed in glass	1.0 Maximum, According to Mounting	1.0 Minimum, According to Mounting	—

1.3.5 Dissipation Constant

The dissipation constant is the reciprocal of the power sensitivity and is expressed as the power input necessary to raise the temperature of the element by 1 degree centigrade (see Table 1).

1.3.6 Voltage/Current Curves and Maximum Voltage Factor

The voltage and current characteristics are of considerable importance (see Figure 2). As the current through a thermistor is increased, the voltage at first rises rapidly and soon attains a maximum value E_{max}. Thereafter, there is a rapid drop in voltage with further increase in current. As the current increases, the rate of fall of voltage decreases and may, under suitable conditions, remain fairly constant over a wide range of current.

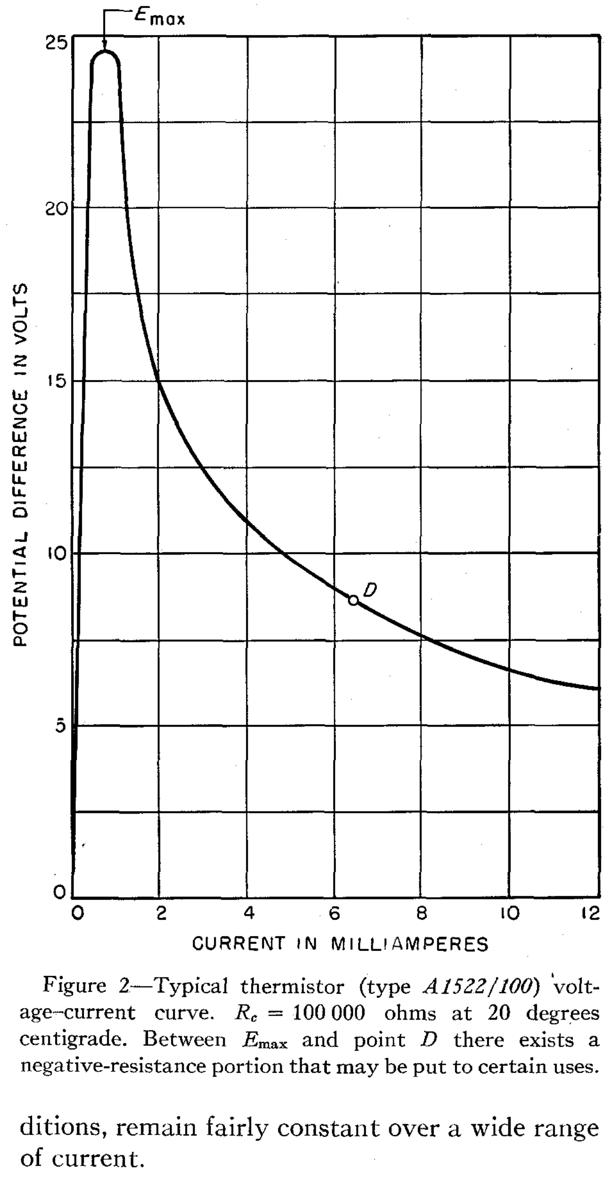

Figure 2—Typical thermistor (type *A1522/100*) voltage–current curve. $R_c = 100\,000$ ohms at 20 degrees centigrade. Between E_{max} and point D there exists a negative-resistance portion that may be put to certain uses.

It can be shown[2] that an approximate value of E_{max} for any thermistor can be calculated by means of the following formula:

$$E_{max} = (R_c)^{1/2}/K,$$

where R_c is the resistance at the ambient temperature (that is, with no dissipation) and K is known as the maximum voltage factor and is dependent only on the ambient temperature and the type of thermistor. Its value for each type at an ambient temperature of 20 degrees centigrade, is given in Table 1.

Thus for a type-*A5412/100* thermistor, which at an ambient temperature of 20 degrees centigrade has a resistance of 50 000 ohms and a maximum voltage factor of 20, the value of E_{max} is

$$(50\,000)^{1/2}/20 = 11 \text{ volts, approximately.}$$

1.3.7 Time Constant

The time constant of an indirectly heated type is defined as the time taken for the logarithm of the resistance of the thermistor to change by $(e-1)/e$ (that is, 63 per cent of the final change) when the heating power is suddenly applied or removed. It is usual to measure the time constant for a change of heating power which, if applied until equilibrium is reached, will halve (or double) the resistance value. It is found that in the high-vacuum types the time constant for a decrease of heating power is longer than for heating, whereas in the gas-filled types, the time constants (heating and cooling) are more nearly the same, and generally speaking, shorter than for the high-vacuum types. The average time constant of a type-*B* thermistor is of the order of 3 or 4 seconds, while that of the type *L* is of the order of 1 to 2 seconds. The time constant of a directly heated thermistor is measured by reducing the dissipation in the thermistor from a large value to zero because of the difficulty of maintaining a constant dissipation (other than zero) in an element whose resistance is changing. For the same reason, the transient behaviour of a directly heated thermistor will vary from circuit to circuit, depending on the manner in which the dissipation alters with the resistance of the thermistor.

1.4 Stability of Characteristics

After the thermistor has been subjected to the ageing treatment that is a normal part of the manufacturing process, there is every reason to believe that it retains its characteristics unchanged after long usage. Life tests have been carried out both with continuous heating and also with the heating current switched on and off every 20 seconds; the original characteristics are maintained after many hundreds of hours of life test.

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[2] J. A. Becker, C. B. Green, and G. L. Pearson, "Properties and Uses of Thermistors—Thermally Sensitive Resistors," *Electrical Engineering*, volume 66, pages 711–725; November, 1946.

Experience has shown that the life of a thermistor depends upon its correct use in a circuit. The following simple rules have been drawn up as a guide to the operating conditions under which a thermistor will give good service.

A. The potential applied to the thermistor element must not exceed $E_{\max}$ when the circuit series resistance is less than 1.25 per cent of the thermistor cold resistance.

B. The potential applied to the thermistor element must not exceed twice $E_{\max}$ when the circuit series resistance is greater than 1.25 per cent of the thermistor cold resistance.

C. The maximum power that may be continuously dissipated is that indicated by the maker. For periods up to one minute, however, this figure may usually be exceeded by 50 per cent.

2. *Some Applications of Thermistors*

Thermistors have come into very wide use in recent years for a variety of different applications. Their small size, ruggedness, stability, and versatility make them extremely useful circuit elements, capable in many cases of fulfilling a function hardly possible by any other means.[3]

It is not necessary to list here all the possible applications of thermistors, but a very broad outline of the basic uses will be given accompanied in many cases by a fundamental circuit diagram.

2.1 TEMPERATURE MEASUREMENT

The use of thermistors for temperature measurement stems directly from their large temperature coefficient of resistance (approximately 10 times that of platinum), which makes possible very sensitive thermometers. In fact a thermistor thermometer has been described[2] that indicates changes of 0.0005 degree centigrade. The high resistance of a thermistor makes the effects of long leads negligible in most cases and the small thermistor element can be very usefully employed for remote indication, particularly as the indicator can be a relatively inexpensive direct-current meter. Temperature measurements of intravenous blood[4] and supercharger rotors, and the insertion of tiny bead thermistors into journal boxes, gear housings, and power transformers during construction are good examples of this type of use. In radiosonde telemetering systems,[5]

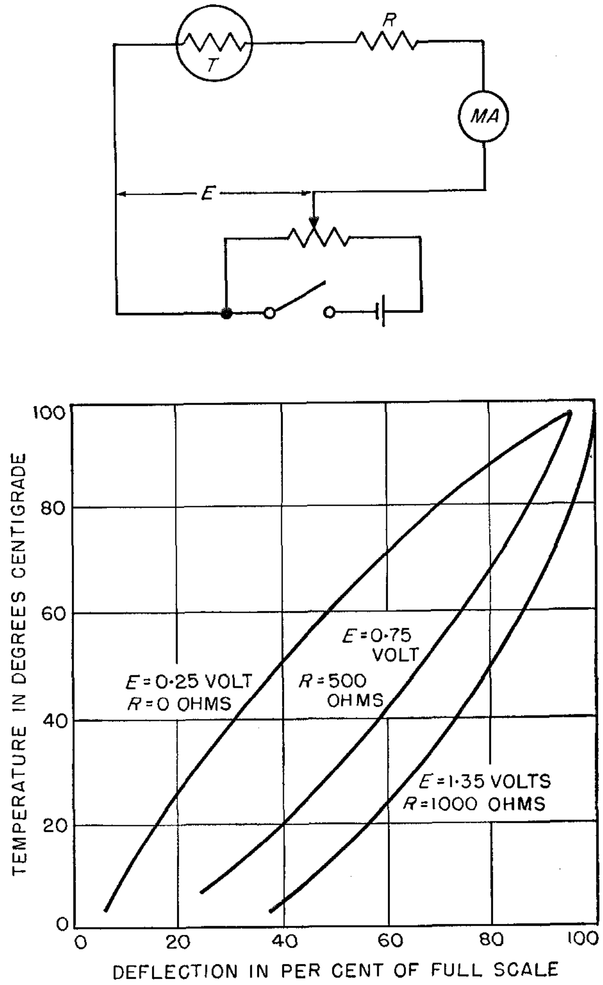

Figure 3—Thermistor thermometer. Type-*A1311/100* thermistor is employed; R_c = 1000 ohms at 20 degrees centigrade. The 1-milliampere meter has a 100-ohm resistance. Characteristic curves are given below.

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[3] J. E. Tweeddale, "Thermistors—They Perform Countless Control Jobs," *Western Electric Oscillator,* pages 3–5, 34–37; December, 1945.

[4] L. F. Drummeter and W. G. Fastie, "A Simple Resistance Thermometer for Blood-Temperature Measurements," *Science,* volume 105, pages 73–75; January 17, 1947.

[5] V. D. Houck, J. R. Cosby, and A. B. Dember, "Radiosonde Telemetering Systems," *Electronics,* volume 10, pages 120–123; May, 1946.

the remote indication is carried out over a radio link. Mention should also be made here of a type of thermistor developed specially for thermometry (type *F* in Table 1), which has a very small mass of thermistor material fused into the glass tip of an evacuated bulb, so that it responds almost instantly to temperature changes.

Figure 3 shows a simple circuit using a thermistor for direct temperature indication. The calibration curves show how the scale linearity and the possible range of temperature measurement can be controlled by changing the battery voltage and the value of the series resistance. For more precise work, a Wheatstone bridge circuit may be used with the thermistor as the unknown resistor. The temperature may be read either directly on the indicator, or by null balance of the bridge; this latter method giving greater accuracy.[6] The fundamental circuit is given in Figure 4. R_1 and R_2 are the ratio arms, *T* is the thermistor, and R_3 is a known resistor, either fixed or variable depending on whether the direct-reading or null-balance method is to be used.

A principal factor to consider in the design is the selection of circuit components and voltage so that the power dissipated in the thermistor itself is too small to produce sufficient self-heating to raise its temperature appreciably above that of its surroundings. The permissible dissipation can be determined from the dissipation constant. The slope of the resistance/temperature characteristic decreases with increasing temperature, so that it is advisable to dissipate the smallest possible amount of power in the thermistor in order to obtain maximum sensitivity. If substantial self-heating is present, the thermistor operates on a non-linear portion of its characteristic (see Figure 1).

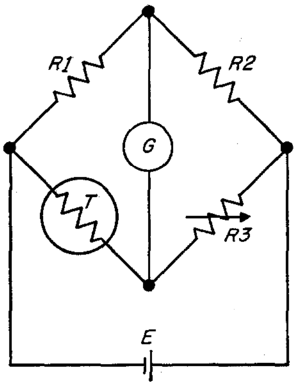

Figure 4—Wheatstone bridge circuit for temperature measurement.

When measuring very small temperature differences, it is sometimes desirable to compensate for ambient temperature variations. R_1 or R_3 may be replaced with a suitable thermistor the characteristic of which is well matched to that of *T*. Any change in ambient temperature then causes an equal change in resistance of both thermistors and the bridge remains in balance.

An object may be ordinarily inaccessible to direct temperature-measuring devices, either through its physical position or because it is in motion, or it may be too hot for contact thermometry. In such cases, its temperature may be determined by permitting radiation from the object to be focussed on a suitable thermistor, as in the noteworthy case of the infra-red bolometer developed during the war to detect the radiation from submarine engine exhausts. This utilised a flake thermistor so sensitive that it would detect the warmth of a human being a quarter of a mile away.[7] Another example is the method used by a manufacturer of tin cans to measure the temperature of the soldered seams while the cans were moving past on a conveyor belt at a rate as high as 350 cans per minute.[8] As it was impossible to make physical contact with the surface being measured because of its rapid motion and inaccessibility, radiant energy was the only energy available as a means of temperature measurement. An optical system was arranged to focus radiant energy from the cans on a flake thermistor and the variation in the amount of energy radiated as the cans moved past the line of sight of the optical system was sensed by the thermistor. The resulting signal voltage, suitably amplified, was presented as a pattern on a cathode-ray oscilloscope.

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[6] E. L. Deeter, "Null Temperature Bridge," *Electronics*, volume 21, pages 180–186; May, 1948.

[7] "Thermal Detectors," *Electronic Industries*, volume 5, pages 87, 116–117; September, 1946.

[8] K. P. Dowell, "Thermistors as Components Open Product Design Horizons," *Electrical Manufacturing*, volume 42, pages 84–91, 190, 192, 194 . . . 216; August, 1948.

2.2 Temperature Control

The same basic requirements apply to temperature control as for temperature measurement, in that the sensitivity should be high and the stability good. In addition, the resistance of the thermistor at the control temperature should be of a value suitable for the circuit used. The circuit shown in Figure 4 may be used with the indicator replaced by a relay that controls the heating or cooling device so that it is switched on or off, according to the requirement, when a predetermined temperature is reached. Normally a valve amplifier is required and the circuit shown in Figure 5 is suitable. The allowable power dissipation in the thermistor elements will be determined by their dissipation constants as for thermistor thermometers. Thermistor bridges have been used for the control of domestic heating systems, refrigeration, and food processing.

To maintain a constant temperature inside a room, a heating system need only balance the heat loss through walls and windows et cetera. There is a time lag in conventional control methods between the actual loss and the increased supply from the heat source to balance the loss. A simple thermistor bridge circuit provides an effective control by making use of two thermistors, either in series or in opposite arms of the bridge, one being mounted close to the heat source and the other outside the house. The bridge is initially balanced to provide the desired room temperature and under these conditions the heat supplied exactly balances the heat losses to the exterior of the house. When the outside temperature falls, the resistance of the outside thermistor rises, unbalancing the bridge. The out-of-balance voltage is used to control the supply of heat, raising the temperature of the heat source and of the second thermistor, until the bridge balance is restored.[8]

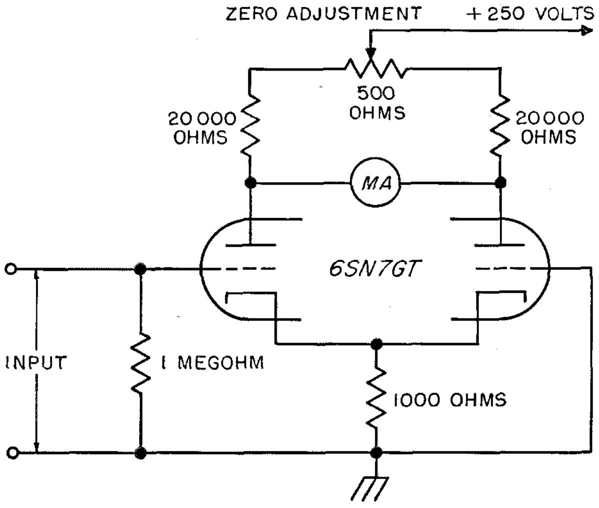

Figure 5—Valve voltmeter for use as indicator with Wheatstone bridge. A 1-milliampere meter is used.

Figure 6 shows a simple arrangement that has been used in a manufacturing process to control the temperature of a heating oven to within ± ½ degree centigrade of any desired temperature in the region of 200 degrees centigrade. The variable resistor is a vernier adjustment to set the grid bias of the thyratron and hence the temperature at which control is to be effected; it may be fitted with a dial calibrated in degrees. A battery is used to supply the thermistor circuit to avoid the effects in the grid circuit of an unstable mains voltage.

A similar arrangement, Figure 7, but completely alternating-current operated, has been used[9] to control the temperature of a domestic refrigerator to within ± 1 degree centigrade. At this lower temperature, the thermistor is more sensitive and the effects of mains variation are not serious. The arrangement is very effective, the refrigerator motor starting immediately the door is opened.

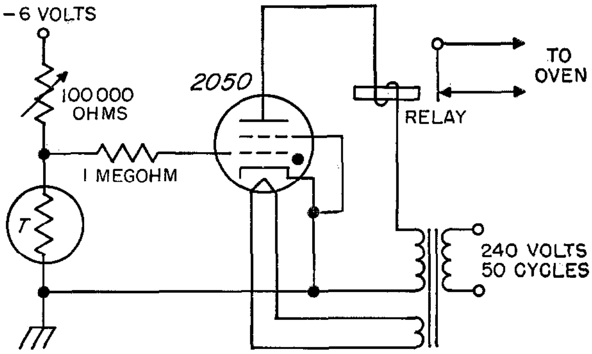

Figure 6—Temperature-control circuit. Type-*F1512/300* thermistor is used; R_c = 100 000 ohms at 20 degrees centigrade.

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[9] A. H. Taylor, "Thermistor Thermostats," *Electronics*, volume 23, pages 154–159; July, 1950.

A more-elaborate circuit has been used[10] in biological research to maintain the temperature of a metal chamber 8 by 8 by 4 inches (20 by 20 by 10 centimetres) at 38 ± 0.1 degrees centigrade while the conduction rate of a small mammalian nerve was being measured. The thermistor is used in a battery-powered bridge circuit followed by a two-stage valve amplifier. A portion of the oven heating element is connected in the plate circuit of the second amplifier stage and variations in plate current due to temperature changes sensed by the thermistor automatically keep the temperature within the prescribed limits. An interesting feature of this arrangement is the continuous variation of current through the heater element in contrast to the on–off switching employed in most control applications.

2.3 Temperature Compensation

The negative temperature coefficient of a thermistor may be used to compensate for the positive temperature coefficient of metallic conductors such as the copper wire in coils or transformers et cetera. Over a small range of temperature the coefficient may be considered linear but in fact it decreases with increasing temperature, as shown in Figure 1.

In its simplest form, the temperature-compensation circuit consists of a metallic resistor and a thermistor in series; any increase in the value of the resistor due to a rise in ambient temperature is compensated for by the thermistor, which is also subject to the same ambient temperature. The temperature coefficient of copper is + 0.4 percent per degree centigrade and that of a typical thermistor may be − 4 percent per degree centigrade, so that if a thermistor having one tenth the resistance of the copper wire is placed in series with it, compensation will be obtained over a narrow range; but such a simple arrangement is not entirely satisfactory as only in rare cases does the law of the thermistor exactly fit that of the resistor. However, by shunting the thermistor with a further resistor, the desired characteristic may be obtained and good compensation may in most cases be obtained over a reasonable temperature range. The resistor used in parallel with the thermistor should have a low temperature coefficient if it is subject to the same ambient temperature change.

Typical applications are the compensation for resistance changes due to temperature rise in the windings of transformers and the temperature compensation of electrical meters.

2.4 Trigger Circuits

The negative-resistance portion of the voltage/current characteristic shown in Figure 2 can be used to provide a trigger action. If sufficient current is passed through the thermistor to make the voltage across it equal to $E_{\max}$, then any small increase of the current or the ambient temperature reduces the thermistor resistance, which increases the current still further, and the resulting increase in current is very considerable, hence the term, trigger. The increase in current may be used to operate a control relay.

This principle has been applied to fire detectors and to sprinkler systems controlled by thermistors. The simplest arrangement is voltage

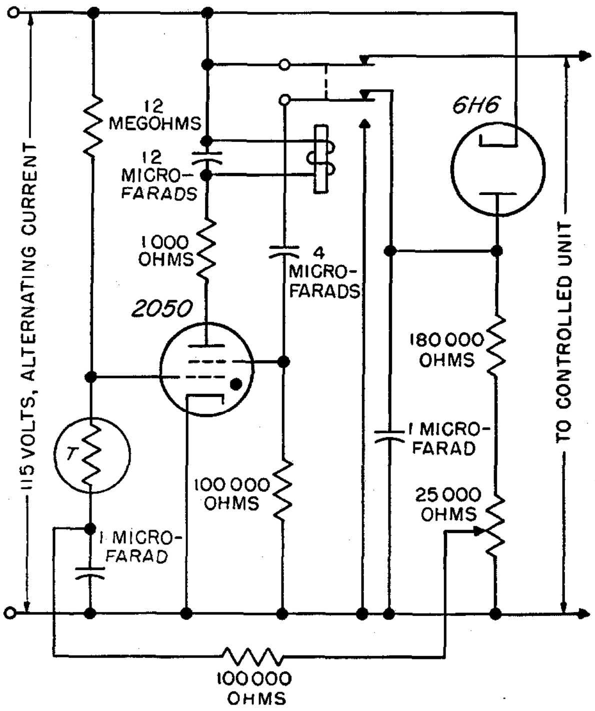

Figure 7—Temperature control circuit.

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[10] C. J. Dickinson, "Thermistor Continuous Temperature Control for Biological Research," *Electronic Engineering*, volume 21, pages 408–409; November, 1949.

source, thermistor, and warning light or bell all connected in series. When the ambient temperature rises beyond normal limits as a result of a fire, the thermistor triggers and operates the warning device. It is noteworthy that prevention of losses and reduction in fire-insurance rates where such devices are used may well repay the cost of installation many times over.

The same principle may be used to provide a control system operated by radio-frequency signals by first adjusting the circuit to the critical point and using the rectified radio-frequency signal to trigger the circuit. By careful adjustment of the circuit, fractions of a milliwatt may be detected and used to operate a relay.

2.5 Timing Devices

The trigger characteristic previously referred to may be used to provide a time delay switch, and delay times from a few milliseconds to many minutes are possible.

A thermistor is wired in series with a relay and, until the thermistor reaches the critical temperature, the current rises slowly. At the temperature corresponding to $E_{\max}$, a large increase in current occurs, which operates the relay. A certain amount of adjustment of this delay time may be obtained by means of a variable resistor in series with the circuit, as well as by choice of the time constant of the thermistor.

In the same way, a relay may be made to open a certain time after closing by connecting a thermistor in parallel with it, and this application may be useful for operations requiring controlled time such as, for example, photographic exposure, although it should not be used where very precise limits on the timing interval are required unless compensation for ambient temperature changes is introduced.

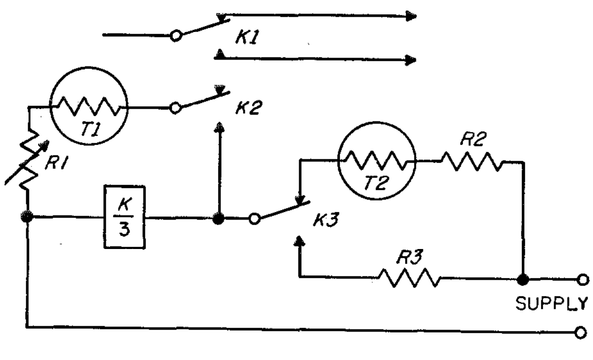

Figure 8—Switching circuit for timing device.

By combining the series and parallel arrangement, a continuously operating timing switch may be constructed. In Figure 8, the relay *K/3* is supplied by the resistor R_2 and the thermistor T_2; by suitable adjustment of R_2, the relay may be made to operate after a predetermined time. When the relay pulls up, K_2 and K_3 operate, removing T_2 from the circuit and placing T_1 and R_1 in parallel; the hold-in current of the relay is maintained by R_3. When T_1 reaches the critical temperature, its resistance rapidly falls, shunting the relay and allowing it to de-energize, and the supply is connected again via T_2 and R_2. Thermistor T_2 has cooled in the meantime, and so the cycle repeats.

2.6 Surge Suppression

Thermistors may be used for surge suppression by connecting them in series with the device to be protected; their high initial resistance prevents any short-duration surge from operating the device, whereas the operating voltage, although of much lower amplitude, is applied long enough to cause heating of the thermistor and consequent reduction of resistance to the point where the device operates. An example is the suppression of the surge of heater current in an alternating-or-direct-current radio receiver when it is switched on. The valve heaters in a receiver of this type are connected in series and supplied from the power mains through a voltage-dropping resistor. The cold resistance of the heaters is as low as a quarter of the hot resistance, and the large switching surge is apt to cause short valve life. A dial lamp is often used in series with the heaters and, due to the fact that it operates at a higher temperature than the heaters, the effect of the surge is even more serious in this case. The inclusion of a thermistor of 2000 to 3000 ohms cold resistance in series with the circuit reduces the surge to less than 10 per cent overload, and when it is hot, the voltage drop across the thermistor will be quite low. The same principles can be applied to reducing the starting current of small electric motors.

A further possibility is the protection of tungsten-filament projection lamps and lamps used for lighting in photographic studios, which have an inherent short life mainly due to the switch-on surge; suitable types of thermistors are being developed for this purpose.

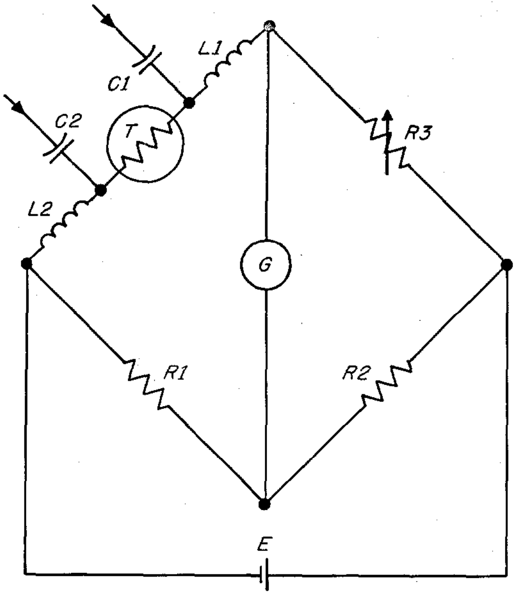

Figure 9—Power measurement circuit.

2.7 POWER MEASUREMENT

The resistance of a particular thermistor depends only on its temperature and, if the ambient temperature remains constant, its resistance is a function solely of the power dissipated in it. A thermistor thus provides a valuable means of power measurement, applicable to any waveform and any frequency up to 10 000 megacycles per second with types at present available commercially. The particular type chosen depends on the frequency.

For relatively low frequencies such as the measurement of the random noise power from an amplifier, an indirectly heated type is the most convenient and can be previously calibrated on direct current or 50-cycle-per-second alternating current. The heater coil is usually about 100 ohms and presents a constant load impedance; the resistance of the bead is then a measure of the root-mean-square power in the heater coil. A convenient method is to heat the thermistor first with direct current to a certain temperature and thus to a certain resistance. When the energy to be measured is then applied to it, its temperature rises and its resistance drops. The direct-current energy is then reduced until the resistance and temperature regain their original values and since the heat dissipation depends only on the temperature, the reduction of the direct-current energy will be equal to the alternating-current energy supplied.

At high frequencies, the indirectly heated types cannot be used on account of the heater inductance, and it is more convenient to use the directly heated types in a bolometer bridge as shown in Figure 9. The thermistor is included in a normal bridge circuit but is isolated for radio frequencies by two radio-frequency chokes L_1 and L_2. The thermistor is calibrated first on direct current and a curve drawn of input power versus resistance. Radio-frequency power is then applied to the thermistor by means of R_3. The resistance value thus measured may be transferred to a power measurement from the curve.

2.8 AMPLITUDE STABILIZATION

For an oscillator to have good frequency stability and waveform it is necessary to provide some means of amplitude stabilization other than that of the curvature of the characteristics of the oscillator valves: Two methods are in general use; namely, the use of a diode rectifier feeding a variable bias voltage to one or more of the oscillator valves or a tungsten-filament lamp connected in the feedback circuit. The use of a thermistor leads to a simpler circuit than the diode rectifier, and it consumes much less power than the tungsten-filament lamp, besides working at a more convenient impedance level.[11] A typical method is shown in Figure 10, where any increase in oscillator output causes an increase in power dissipation in the thermistor, thus lowering its resistance, which reduces the input to the grid of *V1* and tends to restore conditions to normal.

Thermistors have been extensively used to stabilize the levels of carrier- and audio-frequency amplifiers in wire-communication systems.[12] Negative-feedback-type amplifiers are particularly suitable for this application and a simplified circuit is shown in Figure 11. An increase in output level causes more power to be dissipated in the thermistor, its resistance decreases and thus the amount of negative feedback increases until the level is restored to the original value. An indirectly heated thermistor may also be used in much the same way, the heater being connected to the amplifier output.

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[11] L. A. Meacham, "Bridge Stabilized Oscillator," *Proceedings of the IRE*, volume 26, pages 1278–1294; October, 1938.

[12] J. H. Bollman, "Pilot-Channel Regulator for the K1 Carrier System," *Bell Laboratories Record*, volume 20, pages 258–262; June, 1942.

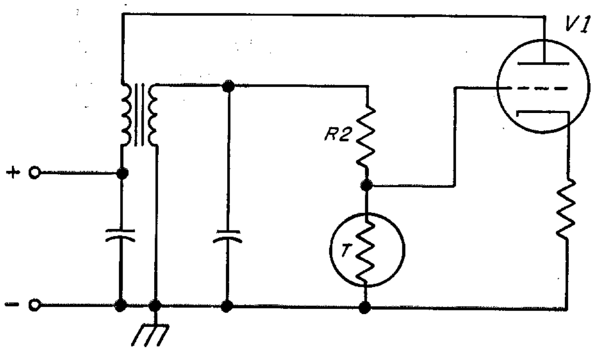

Figure 10—Oscillator amplitude-stabilization circuit.

2.9 LOW-FREQUENCY OSCILLATORS

An oscillator capable of generating small outputs at low frequencies may be constructed making use of the negative-resistance characteristic shown in Figure 2 between $E_{\max}$ and point D. By suitably biasing the thermistor so that it is operating in this region, the negative resistance may be utilized to compensate for the positive resistance of the inductors and if sufficiently negative will permit oscillation, provided the frequency is low enough to allow the thermistor to follow the cycles thermally. Experimental thermistors of extremely short time constant have been developed in the laboratory to cover the audible range and future developments may even extend the range, although these types are not yet commercially available.

It is also possible to use a thermistor as a relaxation oscillator in a resistance–capacitance circuit similar to a thyratron sawtooth oscillator by using the trigger properties mentioned previously.

2.10 VARIABLE RESISTORS

Indirectly heated thermistors are very suitable for use as variable resistors of low wattage rating, and a smooth control over a range of 100 to 1 is readily obtainable.[13] There is a slight time lag due to the thermal inertia of the thermistor but this may not always be of importance. The heater and the thermistor are isolated from each other electrically and therefore can be connected into separate circuits. Useful possibilities are suggested when it is considered that this provides a resistor whose value is proportional to a voltage or current without the need for an intermediate electronic/magnetic transfer device.

This principle may also be used for the remote control of resistors from a considerable distance. For example, an indirectly heated thermistor with a heater of 100 ohms resistance may have an element of several hundred thousand ohms and this may be varied by controlling the current in the 100-ohm heater circuit, which provides no problems over long distances.

2.11 MEASUREMENT OF VELOCITY OF FLUIDS

The velocity of a stream of fluid may be measured[14] by its cooling effect on thermistors over a range of approximately 0.4 to 4 feet per minute. A bridge circuit is used with a compensating thermistor exposed to the same ambient temperature as the measuring thermistor, but screened from the cooling blast, in the same way

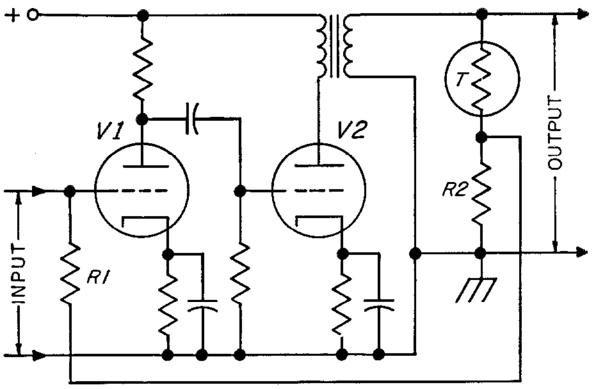

Figure 11—Negative-feedback circuit.

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[13] C. C. Roloff, "Thermo-Variable Resistors; Glass-Bulb Urdox Resistor," *Electrical Review*, volume 140, pages 315–316; February 14, 1947.

[14] "Use of Thermistors to Measure Temperature and Wind Velocity," Office of Scientific Research and Development Report 4699, Department of Commerce, Washington 25, District of Colombia; Bureau of Publication Report PB 13102, Appendix C.

as the pitot and static tubes are used on airspeed indicators to compensate for changes in atmospheric pressure. A known power is dissipated in the thermistor and its resistance measured, after which it is exposed to the cooling blast. The greater the rate of flow, the greater the cooling and the higher will be the resistance. The power dissipated is then increased until the resistance is the same as it was without the cooling blast and this difference in power is converted to a temperature change and used to calculate the rate of flow by using the formulae below. These are reasonably accurate for air and water at an ambient temperature of 20 degrees centigrade, provided there is good thermal contact between the thermistor and the medium.

For air cooling,

$$\text{Cubic feet per minute} = \frac{1.75 \times W}{T},$$

for water cooling,

$$\text{Gallons per minute} = \frac{3.15 \times W}{\Delta T \times 10^3},$$

where W is the power difference in watts and ΔT is the temperature change in degrees centigrade.

Micro-meteorology, or the study of the variation of air temperature, humidity, and velocity over very small distances of the order of a foot or so, is an ideal field for the use of thermistors in this manner.[15]

2.12 VACUUM MANOMETERS

The thermal conductance of the surrounding medium has an effect on the relation between the power dissipated by a thermistor and its temperature, and this fact enables it to be used for vacuum measurement and control purposes. In these cases, the thermistor is really being used as a thermal conductance meter,[16] because in a vacuum system the thermal conductivity of the gas is a function of pressure for any particular gas or mixture of gases. When a constant power is dissipated in the thermistor its temperature and hence its resistance is a measure of the gas pressure. The best results are obtained by using a bridge circuit and a measuring range of 10^{-5} to 10 millimetres of mercury is practicable.

Thermistor manometers are particularly useful where the gas used is capable of being decomposed thermally, and the rise in temperature of the thermistor can be limited, if necessary, to no more than 30 degrees centigrade above ambient temperature.

Thermistors have been used[17] in this way to control the temperature of a vacuum furnace used for the manufacture of radio valves. A relay and an indicating meter, energized by thermistor manometers, provide automatic switching-on of tungsten heaters at a predetermined vacuum level and turns them off again if that level is later exceeded, thus safeguarding the heaters against burn-out due to excessive gas pressure.

3. *Acknowledgments*

The material for this paper has been compiled from a variety of sources, and particular thanks are due to Standard Telephones and Cables Pty. Limited, Sydney, for making available many laboratory reports on the subject.

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Recent Telecommunication Development

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Germanium Positive-Gap Diode: New Tool for Pulse Techniques*

By A. H. REEVES and R. B. W. COOKE

Standard Telecommunication Laboratories, Limited; London, England

CERTAIN electroforming processes produce contact-type germanium diodes that exhibit negative-resistance properties. Such diodes may be effectively used to generate pulses corresponding to 20 megacycles per second with rise and fall times not greater than 0·01 microsecond. They may also be utilized as aperiodic double-stability devices, self-restoring trigger circuits, and binary counters.

• • •

1. *History*

The origin of the germanium positive-gap diode lies in a series of experimental researches conducted during 1950 and 1951 into the properties of normal germanium diodes. During electroforming tests that were part of these studies, there was occasionally observed a diode having an unusual property; a negative-slope resistance in a certain part of the forward (conducting) region of the diode characteristic. The possible practical importance of this discovery was at once realized; further studies were undertaken to find the best way of making such unusual diodes reproducibly and with the most desirable characteristics.

The probable importance of such a diode can be seen from the following considerations. For certain telephone circuits of both long and short distances and also for electronic switching in the telephone field, many engineers now consider it to be economically desirable to use up to 1000 speech channels or more on a single transmission path (metallic or radio) with one of the now-existing constant-amplitude pulse systems such as pulse-time or pulse-code modulation. To achieve this on pulse-time modulation would require a pulse rate of at least 6 megacycles per second and on 7-digit pulse-code modulation, 42 megacycles per second. To obtain the usual advantages of freedom from noise and crosstalk with either pulse-time or pulse-code modulation, the rise and fall times of the pulses would both have to be not more than about 0·01 microsecond.

The terminal equipment would have to produce and receive such very narrow pulses at the above high repetition rates and cheap repeaters would also have to be made to handle them. With pulse-code modulation, the terminal must also be able to code and decode pulses of this waveform. For switching, suitable distributor devices or networks would also have to be designed. The basic tools for nearly all these purposes are: (*A*) a nearly aperiodic pulse producer working at a repetition frequency up to about 50 megacycles per second and giving rates of rise and fall of the order of 0·01 microsecond; and (*B*) a similarly aperiodic double-stability device, or binary counter, working to limits of this same order. Owing to their relatively large time-constants, these requirements cannot be met by any existing hard-vacuum or gas-filled valves; some new tool altogether is needed.

A germanium diode with a negative-slope resistance in the conducting direction, however, introduced a completely new range of possibilities, for it was known that the self-capacitances of such diodes could be reduced to at least the same order as that of vacuum valves; while in the conducting direction the diode resistances could be much less than that of the usual valves. The negative-resistance slope, of course, gave the possibility of triggering action.

Tests on samples of positive-gap germanium diodes showed immediate promise of greatly reduced triggering time-constants as compared with valves. It was not long before some diodes were produced on a laboratory scale that nearly met the needed limits of pulse repetition frequency and completely met the needed rates of pulse rise and fall that have been quoted.

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* Reprinted from *L'Onde Electrique*, volume 34, pages 32–37; January, 1954.

2. Construction and Characteristics

A whisker, preferably of silver containing a small quantity of *N*-type impurity such as arsenic, is placed in contact with a germanium surface of *N*-type bulk material prepared by usual methods. Electroforming is then carried out with an unusually high current of the order of 1 to 5 amperes, normally in several bursts of a few seconds each, until the desired voltage–current characterisitic is observed on a cathode-ray-tube monitor with a 50-cycle-per-second sweep. The electroforming current may be either alternating or direct current.

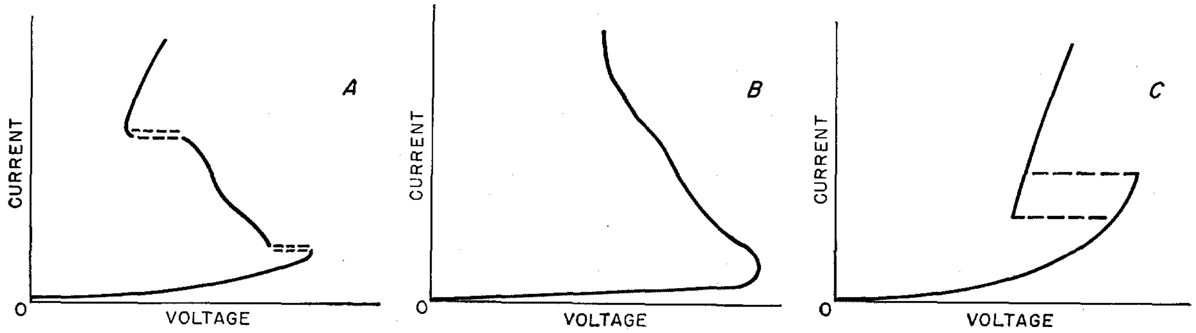

Figure 1—Several types of current–voltage characteristics that may be observed after the electroforming process. The characteristic at *C* indicates a usable positive-gap diode.

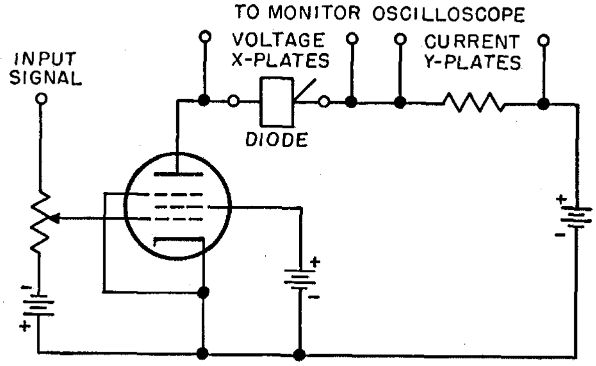

Figure 2—Constant-current source for observing current–voltage characteristics.

The desired shape of the forward (conducting) characteristic explains the name positive gap, for after correct forming there is a gap in the monitor picture, indicating at a certain voltage a double stability in the conducting direction due to a negative-resistance slope. The voltage difference on triggering from one stable position to the other is usually of the order of several volts. The corresponding current change can be 40 milliamperes or more.

Figure 1 shows some typical characteristics, obtained in this case from a constant-current feed derived from a pentode valve, as indicated in Figure 2.

Figure 3 shows the conditions for double stability of a positive-gap diode having the typical characteristic *OABC*, as traced at constant current in a simple series circuit with a battery *E* and a load resistance *R*. Double stability is obtained only if the load line cuts the curve *OABC* three times, as does the straight line *FG*. (The third intersection at *K* represents, of course, an unstable equilibrium point between the two stable points *L* and *M*.) The criterion for this is that the angle θ shall be greater than α, where cot θ gives the load resistance *R* and *HJ*, the tangent to curve *OABC* at point *K*, must therefore not exceed a limiting value. It is also

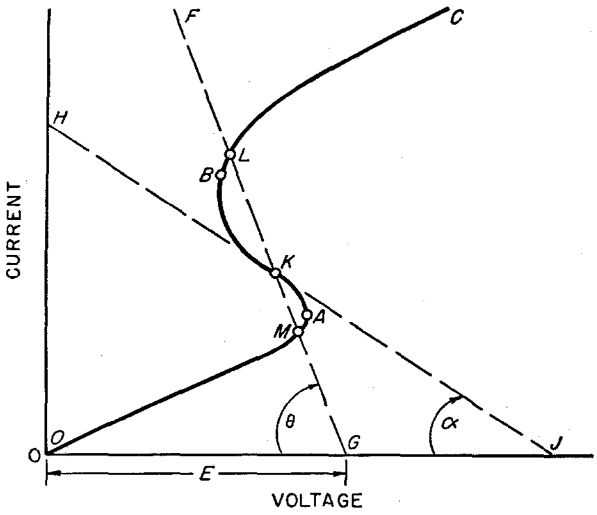

Figure 3—Double-stability characteristic.

necessary that there shall be a suitable value of E for the three intersections to be obtained.

Another typical trace is shown in Figure 4. In this case the trace line is missing between points A and B, showing that even at substantially infinite load slope resistance (from the pentode feed), double stability is still present, the gap AB representing a sharp triggering during the 50-cycle-per-second sweep. It is to be noted that the current value at B is lower than that at A. It will be evident from the two load lines FG and HJ that this type of characteristic can show double stability at all values of load resistance from zero to infinity.

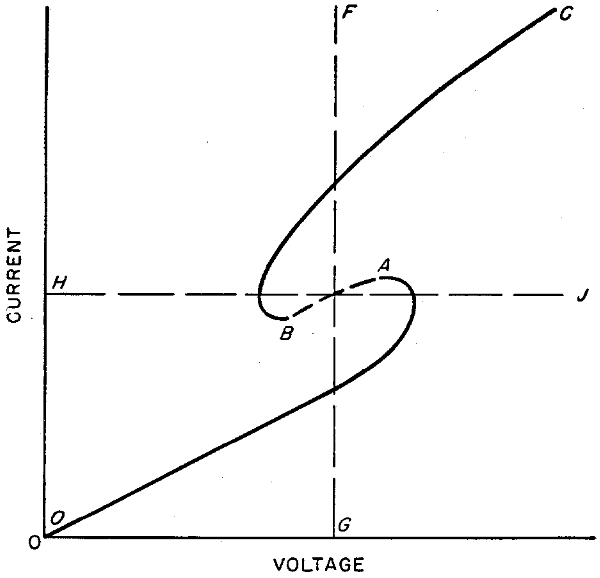

Figure 4—Double stability with zero-to-infinite load-resistance variation.

A characteristic of the shape $OABC$ (Figure 4) can be explained by assuming that there are at least two paths in parallel from the whisker to the bulk germanium, one showing a voltage–current characteristic similar to that of Figure 3 and the other having substantially a pure positive-slope resistance of a suitable value.

3. *Diode as Narrow-Pulse Generator*

Although by special devices pulses have been obtained for laboratory or test purposes of duration times of 0·001 microsecond or less, the apparatus involved is usually somewhat complex. A very simple and cheap method of obtaining pulses at up to, say, 20 megacycles per second repetition frequency with a rise or fall time not exceeding about 0·01 microsecond is clearly of some practical interest for many applications. The positive-gap diode provides a convenient solution to this problem.

One very simple circuit approximately fulfilling these requirements is shown in Figure 5. The whisker of the positive-gap diode D is connected to ground via the small resistor *R1*, the battery, and the secondary winding of the transformer. The base of the diode is grounded via a load impedance shown here as an inductance L shunted by its self-capacitance and the resistance *R2*. The output pulses pass through capacitor *C2*. Capacitor *C1* is very small so as to by-pass only the very-high-frequency components.

Assuming that pulses are required at a steady repetition frequency, a sine wave is applied at this required frequency to the input terminals. The battery voltage and resistor *R1* are then adjusted so that the diode is swept through its triggering point twice in each period of the applied sine wave—once with rising current and once with falling current. With critical damping, two very sharp pulses are thus obtained across the load impedance for each sine-wave period, one positive and one negative. With a damping less than critical, a damped train will result at the output terminal—an adjustment that is sometimes useful. (See for example section 6, describing the use of this diode as a binary counter.)

With the circuit of Figure 5, the pulse rise and fall times could be estimated fairly accurately by the amount of shock excitation produced at various different natural frequencies of the tuned

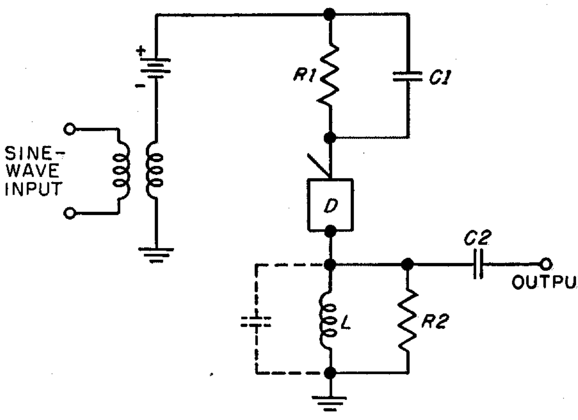

Figure 5—Repetitive-pulse generator.

load circuit. On some samples of diode the time of pulse rise was about 0·0065 microsecond and the time of fall about 0·016 microsecond. This difference in rise and fall time, put to a practical use in the binary counter circuit of section 6, can be explained in terms of the higher internal impedance of the diode in the off stability position as compared with that in the on position. A pulse-repetition frequency of up to 20 megacycles per second was obtained from a number of diode samples. A fairly typical peak output was found to be several volts at about 20 milliamperes, corresponding to a pulse-generator internal impedance of the order of 100 ohms.

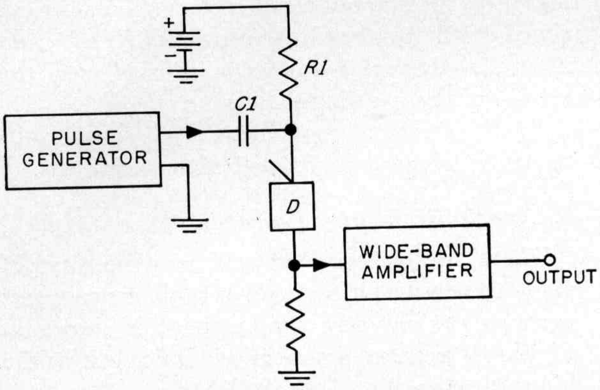

Figure 6—Aperiodic double-stability circuit.

Pulse generators based on positive-gap germanium diodes in the circuit of Figure 5 are now in use at Standard Telecommunication Laboratories; they have been found to be very practical and simple tools for experimental work.

4. *Aperiodic Double-Stability Device*

Figure 6 shows the first circuit using the positive-gap diode as an aperiodic double-stability device. The diode *D* was connected to ground at the whisker end via the small impedance *R1* and battery and at the base end via a small resistance of 80 ohms. A pulse generator produces alternating positive and negative pulses at repetition frequencies up to 20 megacycles per second. This function was performed by distorting a sine wave. The generator was coupled to the whisker of the diode via the small capacitor *C1* and the output from the diode was observed on a cathode-ray tube at the output terminal, after raising the level sufficiently in the wide-band amplifier.

On some diode samples the results showed substantially aperiodic triggering backwards and forwards by the input pulses at frequencies up to 20 megacycles per second, the limiting value of the test equipment. On other diode samples, the triggering ceased at somewhat lower repetition frequencies.

To test the diodes at frequencies higher than 20 megacycles per second, the arrangement shown in Figure 7 was used. The base of the diode *D* was connected directly to ground, while the whisker was grounded via the resistor *R1* and the battery as before. A sine-wave oscillator, adjustable up to about 50 megacycles per second, was arranged to apply several peak volts across the diode via a transmission line and impedance-matching device. The output of the positive-gap diode was observed on a special high-frequency type of cathode-ray tube at the output terminal. A wide-band amplifier was not needed because the diode was chosen to have a large enough voltage difference between the two stability points to make further amplification unnecessary.

When the sine-wave voltage applied to the diode was insufficient to cause triggering, the output trace on the cathode-ray tube was nearly sinusoidal, as shown in Figure 8A. When diode triggering occurred, the kinks in Figure 8B were observed. The tuned circuit associated with the diode was too highly damped to suppress the harmonics represented by these kinks.

By this means two diode samples showed substantially aperiodic triggering, backward and

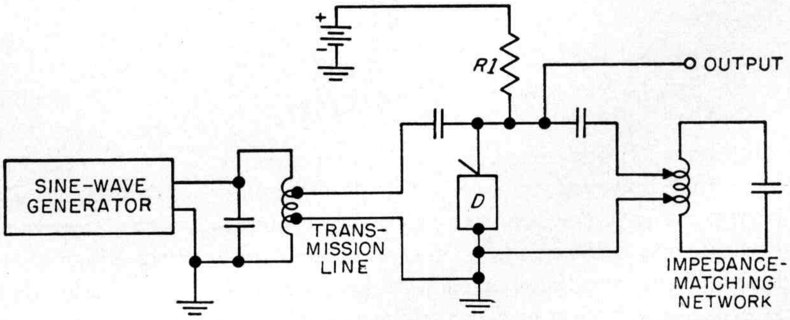

Figure 7—Double-stability testing circuit for use up to 50 megacycles per second.

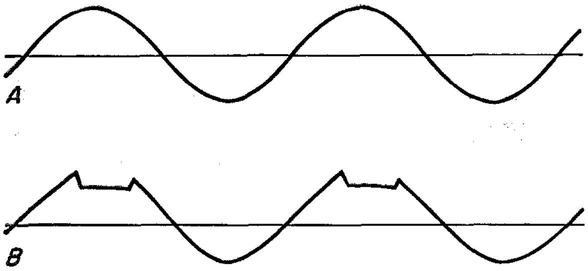

Figure 8—Waveforms on monitor for Figure 7. Kinks indicate double-stability conditions.

forward, at repetition rates of up to 43 megacycles per second.

5. Self-Restoring Trigger Circuit

As shown in Figure 9, the positive-gap diode with grounded base is connected to ground at the whisker end via resistor *R1* and the battery. Unidirectional input pulses are applied across the diode in some suitable manner; for example by means of a pulse generator and transformer. If capacitor *C1* be added across *R1*, suitable values of *C1* and *R1* will cause a relaxation oscillation to be produced—providing that the battery and resistance *R1* are such as to bring the diode into a suitable negative-resistance region on its characteristic. If the voltage across the diode is changed slightly to bring the latter just outside the negative-resistance portion of the curve, oscillation will then cease. If the input pulses are arranged, however, to sweep the diode voltage once more into the negative-resistance region at their peaks, half-periods of the relaxation oscillation can be produced at each such input pulse, alternate oscillation half-periods taking place independently of input pulses and restoring the initial circuit conditions after an adjustable time interval depending chiefly on the value of the time-constant of *C1–R1*.

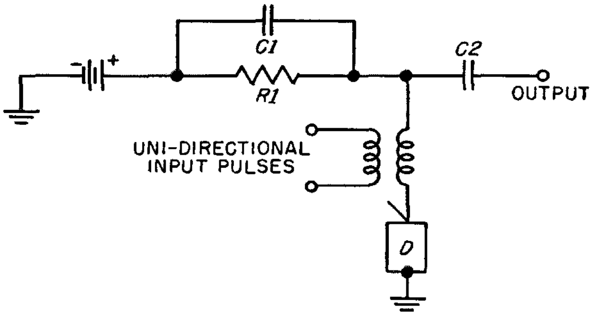

Figure 9—Self-restoring trigger circuit.

There is thus a half-period of free relaxation oscillation timed by each input pulse; this is followed by the start of a self-restoring half-period, timed chiefly by the time-constant of *R1–C1*; and after this there is a quiescent interval until the arrival of the next input pulse.

In this way a convenient and stable self-restoring trigger circuit has been constructed.

6. Aperiodic Binary Counter

There are, of course, several possible ways of using the positive-gap diode as a binary counter; perhaps the simplest arrangement is shown in Figure 10, where only one diode is needed for the binary feature.

The first positive-gap diode *D1* shown in the figure is used to convert the original signal wave to be counted (one having a rounded waveform) into the series of sharp pulses that is needed for the binary input; it is substantially the circuit already shown in Figure 5. Positive-gap diode *D1* has its base grounded via a resistor *R2* (100 to about 400 ohms, in typical cases). The whisker of *D1* is connected to ground via the input transformer, resistor *R1*, and the battery. The small capacitor *C1* is used as a by-pass for the sharp pulses produced by triggering. As in the case of Figure 5, therefore, the (assumed) rounded input waveforms from the input transformer, which are the signals to be counted, are converted into very narrow pulses at the output of capacitor *C2*. These narrow pulses then shock-excite the tuned circuit formed by inductance *L* and its effective self-capacitance shown in broken lines; the resonant frequency being in typical examples in the region of 100 megacycles per second. The circuit of diode *D2* is completed by resistor *R3* and the battery, shunted in the 100 megacycles-per-second region and above by the very small capacitor *C3*. The binary-counter output is obtained at the output terminals via capacitor *C4*.

The operation of the circuit will now be explained. In Figure 11, curve *A* is the assumed input waveform. Curve *B* represents the current waveform through diode *D1*. Curve *C* is the derivative of *B*. It will be noted that the negative-going pulses are smaller than the positive pulses; due to the fact that in a positive-gap diode the internal time-constant on triggering from "current-low" to "current-high" is normally considerably less than the time-constant in the reverse sequence—a phenomenon shown in curve *B* as a steeper slope in the rise portion than in the fall portion. By virtue of small capacitor *C2*, the derived curve shown in *C* shock-excites the tuned circuit *L* and its distributed capacitance, giving a waveform represented as curve *D*. There are two damped trains; an initial train *a* due to shock excitation by the (larger) positive pulse and a following train *b* of relatively small amplitude due to the negative pulse, which is of much smaller peak value than the positive pulse.

The circuit constants of diode *D2* are adjusted so that triggering occurs on the first *negative* peak of the train in curve *D*. (Under the assumed initial conditions, triggering cannot occur on the larger positive peak because this peak is of the wrong sign for that purpose.) Re-triggering to the off position on the second positive peak is eliminated by the internal time-constant of the diode; if the resonant frequency of the tuned load circuit is 100 megacycles per second, for example, the interval between the first negative and the second positive peaks is only 0·005 microsecond, much too short for the diode to be triggered back to the current-low condition. The negative peak at the start of the second following train is too small in amplitude to cause triggering of diode *D2*; therefore this diode will remain in the current-high condition until the arrival of the positive peak at the beginning of the next damped train. Then, diode *D2* will be triggered back to its initial current-low condition.

Neglecting ripples due to pulses, the current through diode *D2* will therefore follow curve *E*. It is thus clear that substantially aperiodic binary-counting action can be obtained in this way.

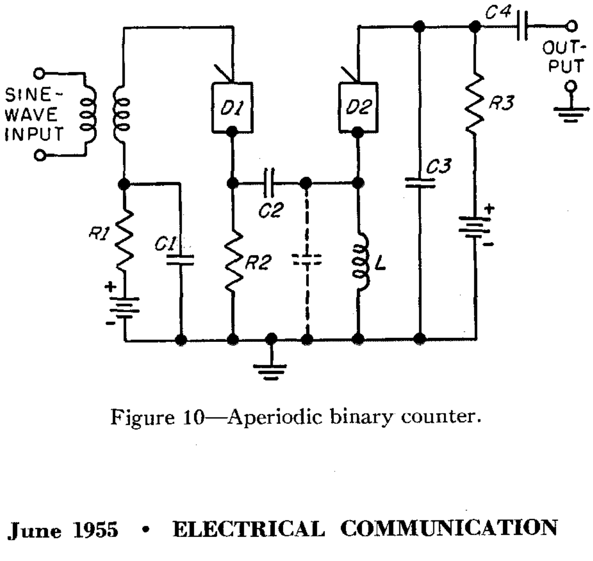

Figure 10—Aperiodic binary counter.

Counting has been achieved on some diode samples at input repetition frequencies up to 15 megacycles per second. A tool such as this, in addition to its uses in telecommunication, should prove valuable in certain types of high-speed electronic calculators, et cetera.

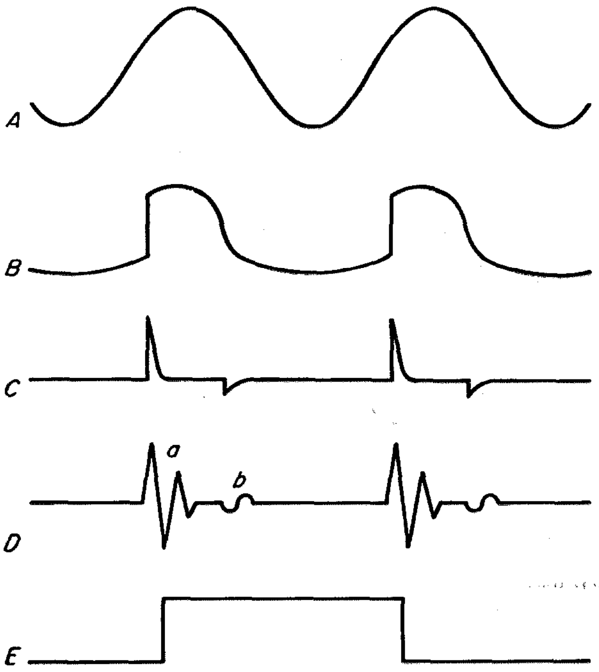

Figure 11—Waveforms for circuit of Figure 10.

7. *Further Uses*

This new form of diode can also be used as a high-frequency oscillator in an *L–C* circuit; a limit of about 300 megacycles per second was obtained from some samples at several milliwatts output power. The diode has also proved to be a successful tool in a very compact super-regenerative receiver in this same frequency range. It should be understood that there is, in fact, a family of such diodes having a range of characteristics and that differing types can be made and employed for various needs. As the purpose of this paper, however, is to discuss only pulse techniques, these further uses will not be described in detail.

Broad-Band Reflectometers at High Frequencies

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A USEFUL instrument in any work with radio frequencies is a device for measuring the phasor sum or difference of two voltages, one being proportional to the current through a load and the other being proportional to the voltage across the load. Such an instrument will measure either the incident or the reflected power in a transmission line. These data may be combined to indicate the standing-wave ratio on the transmission line or the net power being transmitted. Other combinations of the basic "voltage" and "current" voltages can be used to measure magnitude and phase angle of the load impedance or to indicate load resistance or conductance independently of the reactive component.

Well-known forms of this instrument are the reflectometer (or directional coupler) and standing-wave indicator. The variation described here differs primarily in the use of resistive rather than reactive coupling elements, thus eliminating the dependence of sensitivity and insertion loss on frequency. Improved techniques for low radio frequencies permit broad-band use at constant sensitivity from 300 kilocycles per second to 30 megacycles. Applications are extended to provide the sign as well as magnitude of error for servo control; to indicate separately the phase, magnitude, resistance, or conductance of the load; or to measure directly the standing-wave ratio or net delivered power.

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1. *Reflectometer Theory*

The voltage and current at any point in a transmission line may be expressed as the phasor sum of two component waves (see Figure 1) traveling in opposite directions, each having the instantaneous voltage–current relation $e = Z_0 i$, but the currents being taken in opposite directions. The incident wave that is traveling from the source toward the load, therefore, may be expressed at any point by $e_i = Z_0 i_i$ and the reflected wave moving from the load toward the source by $e_r = - Z_0 i_r$.

The phasor relations on which the reflectometer and related wave detectors are based may be obtained from Figure 1.

$$e_t = e_i + e_r, \tag{1}$$

$$Z_0 i_t = e_i - e_r, \tag{2}$$

where

e_t = voltage to ground from a given point on the transmission line
i_t = current at same point on the transmission line
e_i = voltage of incident wave at same point
e_r = voltage of reflected wave at same point
Z_0 = characteristic impedance of the same transmission line.

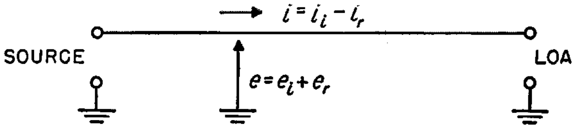

Figure 1—Voltage and current along a transmission line.

From the sum and difference of (1) and (2) may be derived

$$e_i = (e_t + Z_0 i_t)/2, \tag{3}$$

$$e_r = (e_t - Z_0 i_t)/2. \tag{4}$$

The magnitudes of e_i and e_r are constant (except for attentuation) at all points on the line, the voltage standing-wave ratio being

$$\frac{|e_i| + |e_r|}{|e_i| - |e_r|} = \frac{|e_t + Z_0 i_t| + |e_t - Z_0 i_t|}{|e_t + Z_0 i_t| - |e_t - Z_0 i_t|}. \tag{5}$$

Accordingly, if suitably adjusted voltages proportional to line current and voltage are added (or subtracted) vectorially, the rectified result will yield by (3) and (4) a direct voltage proportional to the incident (or reflected) voltage on the line. This is evident from Figures 2 and 3.

1.1 Standing-Wave-Ratio Monitor

By arranging the detecting rectifiers to produce one polarity of direct current for incident-

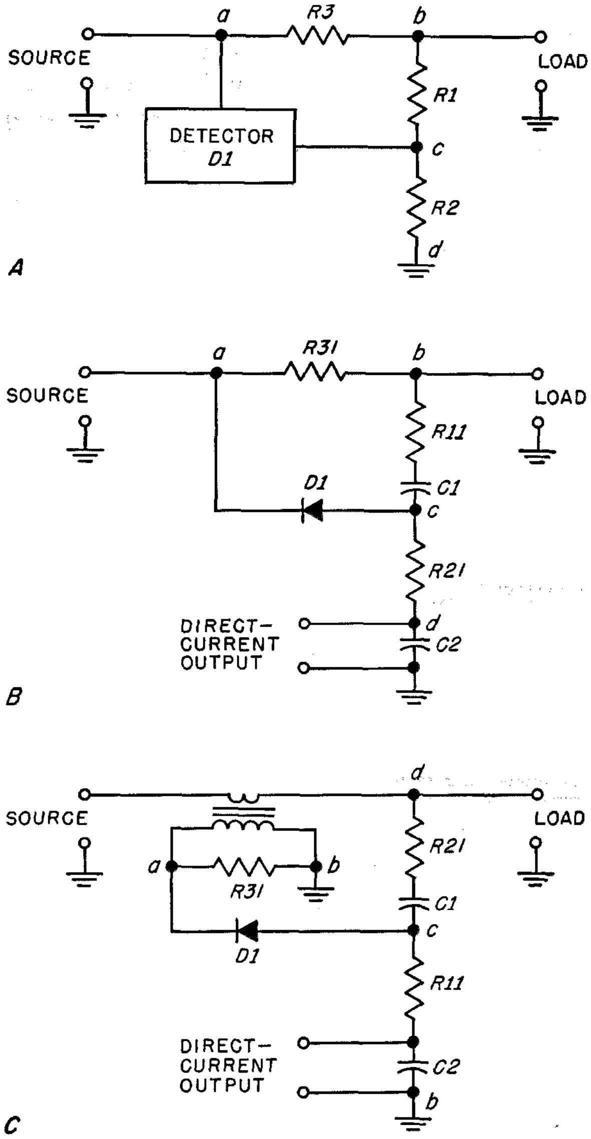

Figure 2—Detection of an incident wave. The rectifiers *D1* are suitable for radio frequencies and the blocking capacitors *C1* and *C2* present a low impedance at the operating frequency. *A* is the simplified schematic circuit, *B* is the current-resistor form, and *C* is the current-transformer type of circuit.

wave output and the opposite polarity for the reflected-wave output, a potentiometer connected between the two can be adjusted to produce a null for any chosen ratio of incident-to-reflected output. This is shown in Figure 4. Such a null circuit gives a definite polarity indication of whether the voltage standing-wave ratio on a line is greater or less than a preselected value and may be used as a limit-bridge type of monitor to accept or reject a given load adjustment.

1.2 Power Meter

If square-law detectors are provided, the difference between the squared values of the incident-wave output and the reflected-wave output of the detectors is a direct measure of total power delivered to the load.

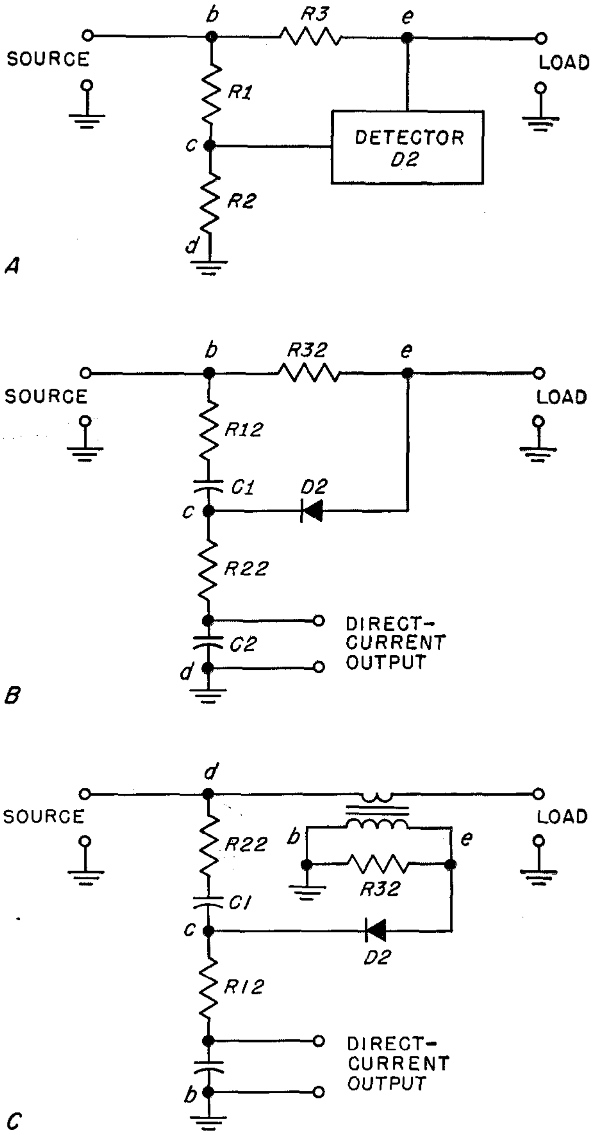

Figure 3—Detection of a reflected wave using a detector and blocking capacitors suited to the operating frequency. *A* is the schematic circuit, *B* is for a current-resistor source of voltage, and *C* is for a current-transformer arrangement.

For the more-usual case of linear detectors (where I and R are obtained, rather than I^2 and R^2), it is also possible to obtain a direct indication of power by using a linear combination of incident and reflected outputs to operate a direct-current wattmeter as is shown in Figure 5.

Since the wattmeter indication is proportional to the product of the currents in the two coils, $I + R$ is supplied to one coil and $I - R$ to the other. The product is obviously $I^2 - R^2$ or the net power delivered to the load.

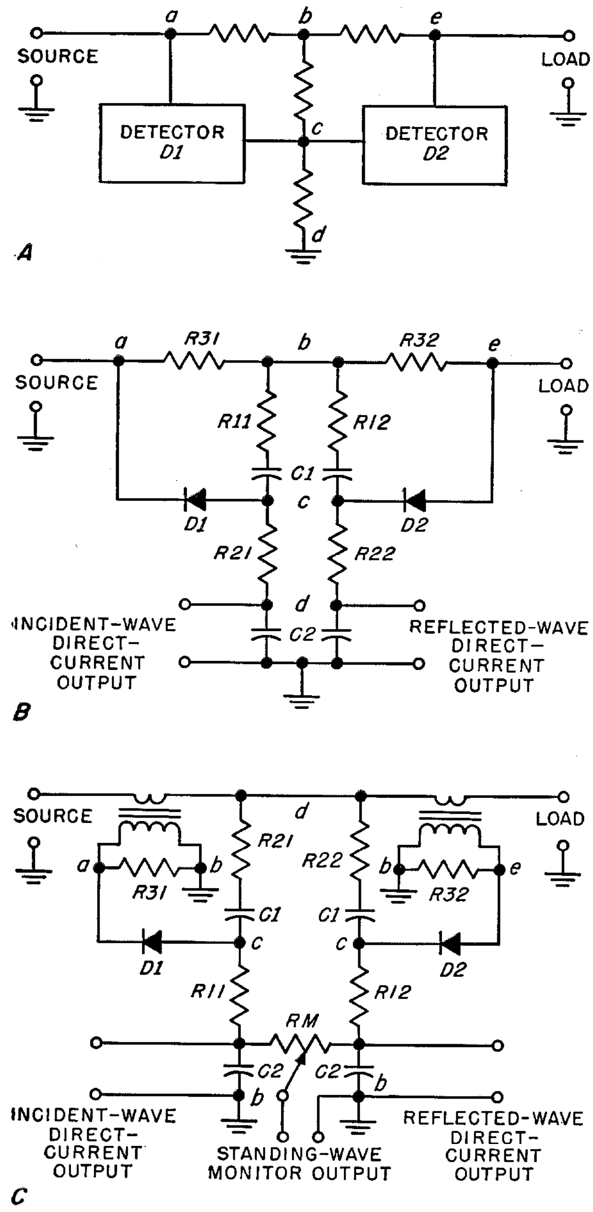
a
b
e
SOURCE
LOAD
DETECTOR D1
DETECTOR D2
c
d
A
R31
R32
R11
R12
C1
D1
D2
R21
R22
C2
INCIDENT-WAVE DIRECT-CURRENT OUTPUT
REFLECTED-WAVE DIRECT-CURRENT OUTPUT
B
RM
STANDING-WAVE MONITOR OUTPUT
C

Figure 4—Combined reflectometers or standing-wave monitor; A being the simplified; B, the current-resistor; and C, the transformer arrangements.

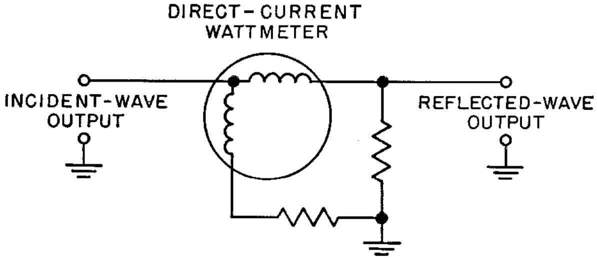

Figure 5—Method of obtaining a direct indication of power.

2. *Measurement of Load Resistance, Conductance, or Phase Angle*

2.1 Resistance Bridge

When the conditions (A), (B), (C), and (D) given below are imposed on the reflected-wave detector, the circuit of Figure 6 is obtained. It responds only to variations in the equivalent series resistance of the load, the null obtained being unaffected by the series reactance. The polarity of the output indicates whether the load series resistive component is greater or less than a chosen value. This circuit is of particular interest if the resistance and reactance of the load are to be adjusted independently.

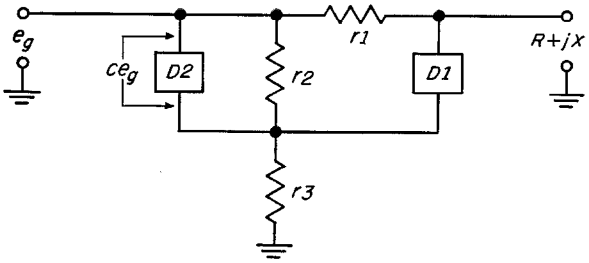

Figure 6—Resistance bridge.

$$e_g = (R + r_1 + jX)i$$

$$c = \frac{1}{2}\frac{r_1}{Z_0 + r_1} \qquad (A)$$

$$D_1 = |ce_g - r_1 i| \qquad (B)$$

$$D_2 = |ce_g| \qquad (C)$$

$$r_2 + r_3 \gg R + r_1. \qquad (D)$$

$$D_1 = \left| \frac{1}{2} \frac{R + r_1 + jX}{Z_0 + r_1} r_1 i - r_1 i \right|$$

$$D_2 = \left| \frac{1}{2} \frac{R + r_1 + jX}{Z_0 + r_1} r_1 i \right|$$

$$D_1 = \frac{r_1 i}{2(Z_0 + r_1)} \left| R + r_1 - 2Z_0 - 2r_1 + jX \right|$$

$$D_2 = \frac{r_1 i}{2(Z_0 + r_1)} \left| R + r_1 + jX \right|$$

$$D_1 - D_2 = \frac{r_1 i}{2(Z_0 + r_1)} \left(|R - r_1 - 2Z_0 + jX| - |R + r_1 + jX| \right).$$

For $R = Z_0$,

$$|Z_0 - r_1 - 2Z_0 + jX| = |Z_0 + r_1 + jX|.$$

$$D_1 - D_2 = 0 \text{ for all values of } X.$$

For $R > Z_0$,

$$|Z_0 + \Delta - r_1 - 2Z_0 + jX| < |Z_0 + \Delta + r_1 + jX|,$$

$$D_1 - D_2 < 0 \text{ for all values of } X.$$

For $R < Z_0$,

$$D_1 - D_2 > 0 \text{ for all values of } X.$$

2.2 Conductance Bridge

A similar modification, shown in Figure 7, permits determination of the load conductance or shunt resistance independent of parallel susceptance or reactance.

$$D_2 = \frac{i}{G_1} = \frac{i_L + G_2 e_L}{G_1} = \frac{e_L}{G_1}(G + G_2 + jB).$$

Let $c = 2\dfrac{G_0 + G_2}{G_1}$,

$$D_1 = ce_L - \frac{i}{G_1}$$

$$= \frac{2G_0 + G_2}{G_1} e_L - \frac{G + G_2 + jB}{G_1} e_L$$

$$= \frac{e_L}{G_1}(2G_0 - G + G_2 - jB).$$

$$D_1 - D_2 = \frac{e_L}{G_1}\left(|2G_0 - G + G_2 - jB| - |G + G_2 + jB| \right)$$

for $G = G_0$.

$$D_1 - D_2 = \frac{e_L}{G_1}\left(|G_0 + G_2 - jB| - |G_0 + G_2 + jB| \right) = 0$$

for all values of B.

For $G > G_0$,

$$D_1 - D_2 = \frac{e_L}{G_1}\left(|G_0 + G_2 - \Delta - jB| - |G_0 + G_2 + \Delta + jB| \right)$$

< 0 for all values of B.

For $G < G_0$,

$$D_1 - D_2 > 0 \text{ for all values of } B.$$

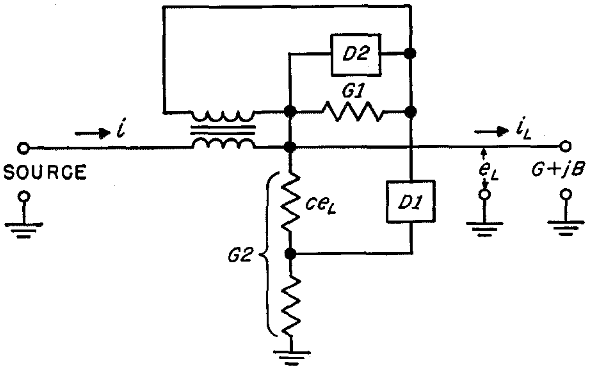

Figure 7—Conductance bridge.

2.3 Phase Bridge

By the addition of suitable phase-shifting networks, shown in Figure 8, a circuit responsive only to the phase angle of the load impedance is obtained. The balance conditions for this bridge are independent of frequency and impedance magnitude as shown by the balance equations. The sensitivity is necessarily somewhat affected, but by proper choice of parameters a practical frequency range of 60 to 1 can be covered with a sensitivity variation of about 4 to 1. For minimum variation with frequency, corresponding reactance and resistance values should be numerically equal (45-degree phase relation) at the geometric mean frequency (that is $X_3 = R_3$, et cetera) and the voltage arm X_5–R_5 should be tapped so as to make $e_{ab} = e_{cf} = e_{de}$ for a matched load.

$$\frac{X_3}{R_3} = \frac{X_4}{R_4} = \frac{X_5}{R_5} = \frac{X'_6}{R'_6} = \tan \alpha$$

$$R_1 = R_2,\ X_3 = X_4, \text{ and } R_3 = R_4.$$

The following approximations are assumed in the discussion.

$e_g = e_i$, $i_g = i$, $R_3 \gg R_1$, and $R_5 \gg R_1$.

$e = e_L = e_g$

$$i = i_L = i_g$$
$$e = i(M \exp [jB])$$
$$e_{ac} = iR_1 = - e_{ec} = - iR_2$$
$$e_{cg} = e$$
$$e_{bc} = (e_{ac} \cos \alpha) \exp [j\alpha]$$
$$e_{dc} = (e_{ec} \cos \alpha) \exp [j\alpha]$$
$$e_{cf} = (e_{cg} \sin \alpha) \exp [j(90^\circ - \alpha)]$$
$$e_{bc} = (iR_1 \cos \alpha) \exp [j\alpha]$$
$$e_{dc} = (iR_1 \cos \alpha) \exp [j\alpha]$$
$$e_{cf} = i(M \sin \alpha) \exp [j(90^\circ + B - \alpha)]$$
$$D_1 = |e_{bf}| = |e_{bc} + e_{cf}| = |i(R_1 \cos \alpha) \exp [j\alpha] + i(M \sin \alpha) \exp [j(90^\circ - \alpha)]|$$
$$D_2 = |e_{df}| = |e_{dc} + e_{cf}| = | - i(R_1 \cos \alpha) \exp [j\alpha] + i(M \sin \alpha) \exp [j(90^\circ + B - \alpha)]|$$
$$D_1 = |i \exp [j\alpha]| \, |R_1 \cos \alpha + jM \sin \alpha \exp [jB]|$$
$$D_2 = |i \exp [j\alpha| \, |]R_1 \cos \alpha - jM \sin \alpha \exp [jB]|$$
$$\exp [jB] = \cos B + j \sin B$$
$$D_1 = |i \exp [j\alpha]| \, |R_1 \cos \alpha - M \sin B \sin \alpha + jM \cos B \sin \alpha|$$
$$D_2 = |i \exp [j\alpha]| \, |R_1 \cos \alpha + M \sin B \sin \alpha - jM \cos B \sin \alpha|$$

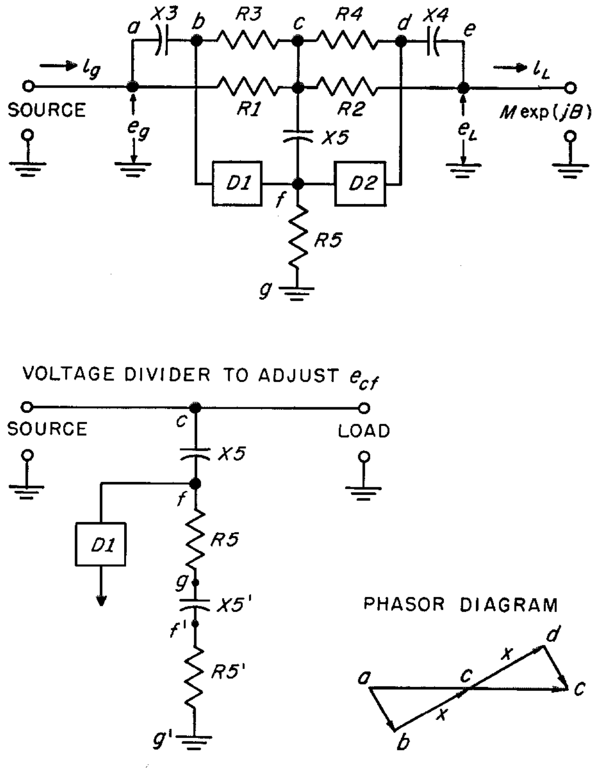

Figure 8—Phase bridge.

For $B = 0$ (load resistive), $\sin B = 0$, and $D_1 - D_2 = 0$ for all values of M.

For $0 < B$ 90 degrees (inductive), $0 < \sin B < 1$, and $D_1 - D_2 < 0$ for all values of M.

For $0 > B > 90$ degrees (capacitive), $0 > \sin B > - 1$, and $D_1 - D_2 > 0$ for all values of M.

As a companion to the phase bridge, a simple modification of Figure 2 with D_1 replaced by separate (oppositely poled) detectors, one across *ab* and the other across *bc*, provides an indication of magnitude of load impedance that is independent of phase angle.

3. Construction

The following methods may be used to obtain the required results at frequencies up to 30 megacycles. At higher frequencies, some developmental work may be necessary.

Resistors appear in all the circuits above and (since phase is measured) must remain resistive over the required frequency range. With average carbon resistors, only values of resistance below 150 ohms can be used in the case of the 2-watt size and values below 250 ohms in the 0.5 watt size. To obtain large values of resistance, it is necessary to use these smaller values in series. Some cancellation of the capacitances occurs because of the lead inductance and in some cases resistors as large as 1000 ohms may be used as building elements. Extreme care is necessary in the latter case because long leads and metallic terminal posts act as capacitances to ground. In general, the leads should be as short as possible with no use of terminal posts except with resistors of less than 100 ohms. If a resistor of 20 000 ohms is needed, the problem is acute and a configuration must be found in which there is a proper balance of (*A*) capacitance from the radio-frequency line to the resistors in the top of the string, (*B*) capacitance among the resistors in the string, (*C*) capacitance between the bottom resistors and ground, and (*D*) distributed capacitance of the string to ground. One successful method of doing this is an arrangement with 20 resistors each of 1000-ohm 0.5-watt rating soldered together in series as shown in Figure 9. The capacitance from the radio-frequency side may be adjusted by varying the position of a wire attached to the radio-frequency line. The capacitance among resistors may be varied by the spacing among them, and the capacitance to ground may be varied by moving the assembly with respect to ground.

There is a lower limit to the value of resistance that may be used as well as an upper limit. This is imposed by the series inductance due to the physical length of a unit between points of attachment and includes lead length. Thus to obtain 1 ohm, it is necessary to use 10-ohm resistors in parallel mounted on metallic sheets to minimize leads.

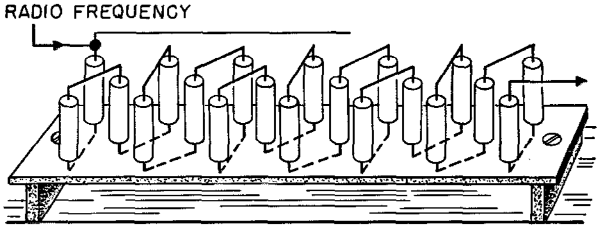

Figure 9—Disposition of 20 resistors to reduce capacitive coupling to a minimum value. A final balance is obtained by dressing with upper rigid lead that is connected to the radio-frequency line.

In those cases in which power is not a problem, there are some carbon-film resistors of comparatively high value that are satisfactory. This situation is not a usual one. Instead, there occurs a need for components that will handle more than the 2-watt capacity of the carbon type. The precious-metal film-type 5-watt size in resistance from 60 ohms to 200 ohms provides for such a requirement. If the power dissipation approaches 50 watts, then the noninductive wire-wound type with calibrated frequency characteristics can be used. The latter may be adjusted conveniently for a pure resistance value by sliding a metallic tube from the radio-frequency end over the outside of the resistor.

The method of determining whether a resistor or resistor assembly is adjusted correctly for little or no phase angle is that of final performance in a circuit. Before being placed in a circuit, the resistors and sections of resistors should be measured on a radio-frequency bridge to insure that the reactive component is less than 2 percent of the magnitude of the resistor. The effect of the final ground configuration should be included in the measurements. When capacitance from the radio-frequency lead to the top of the resistor is used, care should be taken to deduct the added direct capacitance to ground from the measurement.

Final performance is checked by using a flat resistive load and observing the outputs of the detectors for a fixed radio-frequency input power over a frequency range. The output should remain constant and should not vary continuously with frequency. If variations occur, the fault usually may be traced to the divider.

In all the circuits above, one or more series resistors are used to measure current. An extremely convenient substitution for each of these elements is a current transformer in which the secondary is isolated from the primary radio-frequency power and yet produces an output voltage proportional to the current in the load. The circuits of Figures 2C, 3C, 4C, and 7 are then practical. The advantages are increased sensitivity and the convenience of having all direct-current outputs at radio-frequency ground potential. The detectors are in a more-favorable physical position and are less likely to pick up radiated radio-frequency power that causes confusing unbalances.

The construction of the current transformers deserves some description. The best cores for broad-band use are the toroids made up of 1-mil (0.3-millimeter) superpermeability tape. The secondary winding[1] consists of a relatively few turns of flat wire put on by hand. Each turn is placed on top of the previous turn to form a spiral. The insulation is kept at a minimum to reduce leakage. In general, as many secondary turns are used as will produce self-resonance at the center frequency of the range desired. The primary consists of the radio-frequency line passing through the center of the core. A shield

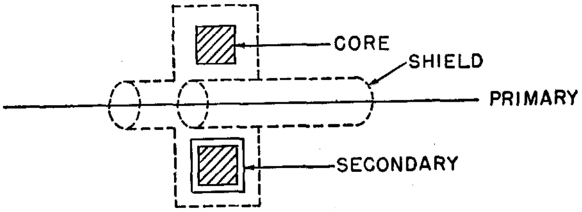

Figure 10—Shielded current transformer.

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[1] D. Maurice and R. H. Minns, "Very-Wide Band Radio-Frequency Transformers," *Wireless Engineer*, volume 24, pages 168–177, 209–216; June and July, 1947.

is used as shown in Figure 10 to reduce leakage and to eliminate electric-field coupling.

The size of the core depends on the power it must dissipate. An average design allows about 1 percent of the primary radio-frequency power to be lost in the core. For 100 watts of primary power, a core of 0.25-inch (6-millimeter) square cross section and 1-inch (25-millimeter) diameter will handle 1 watt. The total open-circuit impedance of a winding is independent of the core size. Large cores require fewer turns than small cores, and the smaller windings produce the same open-circuit impedance because of the decreased magnetic-path length. This, plus the increased efficiency of smaller cores due to a more-equal distribution of core and copper losses, encourages economy of design up to the point where physical size and *leakage reactance* become limitations.

The open-circuit impedances at the two ends of the desired frequency range determine the resistive loading that may be used across the secondary. At the low end, the winding impedance is inductive and at the high end it is capacitive. Resistances across the secondary winding of 10 to 15 percent of the total open-circuit impedance will provide voltages that are constant with frequency and have very little phase shift referred to the current in the primary.

The leakage reactance referred to above may be defined as the flux around any one wire or turn that does not link nearby wires or turns. It is comparable to a series inductance and can usually be kept to a small fraction of its maximum value (the self-inductance of a wire, turn, or winding by itself). Where leakage is large, self-resonance with stray capacitance occurs at a low frequency. The results are a severe drop in open-circuit impedance of the winding at the high-frequency end (the magnitude of $R^2 + X^2$ is about the same on both sides of the center frequency), a drop in output, and undesirable phase shift. These conditions are to be avoided. Where leakage is small, the phase shift produced is sometimes evident, but in most cases can be compensated for in the final adjustment of the voltage divider.

Some practical values of resistances to be used in the above circuits can be given. In Figures 2B and 3B, e_{ab} may be obtained from 1 ohm constructed as previously mentioned. For 100 watts of radio-frequency power, the current into 50 ohms is 1.41 amperes and the voltage across 50 ohms is 70.7 volts making $e_{ab} = 1.41$ volts. To get $e_{bc} = e_{ab} = 1.41$ volts, the ratio of bc/bd is 1.41/70.7 or 1/50. If an 18-ohm resistor is used to obtain e_{bc} then $bd = 50 \times 18 = 900$ ohms. This divider can be made up of one 18-ohm resistor, one 130-ohm resistor, and the rest are 150-ohm units. The total divider is 898 ohms. With 70.7 volts across it, the wattage to be dissipated is 5000/900 or about 55 so that each of the 150-ohm resistors can be 1 watt. The capacitors shown are all for direct-current blocking and radio-frequency bypassing. Any convenient size serving these functions is satisfactory. The detectors may be crystals of the germanium type. A 1-milliampere meter may be used as an indicator with 1000 or 2000 ohms in series. The direct-current return is through the load or generator or through a radio-frequency choke if both the former are open direct-current circuits.

Figures 4B and 4C use two dividers and the resistance of each is double that for Figures 2 and 3 to prevent loading. The direct-current potentiometer at the ground level may be 5000 or 10 000 ohms. In this case, all the current used to generate e_{ab} does not go through the R_{32} resistor because of the shunting effect of the dividers *bd*. Therefore, e_{ab} does not equal e_{be} exactly and some correction is necessary if exact answers are required.

In the current-transformer form, the resistors used to obtain e_{ab} and e_{be} depend on the properties of the cores. A convenient and inexpensive core for the 2- to 26-megacycle range is made of Crolite *70*, a ferrite, and is 1.1 inch (28 millimeters) in outside diameter, 0.6 inch (15 millimeters) in inside diameter, and 0.25 inch (6 millimeters) in thickness. With a 20-turn secondary and a 1-turn primary, the resistance across the secondary may be as large as 300 ohms. This resistance divided by the square of the turns ratio is the resistance reflected in the primary and is 0.75 ohm. For 1.41 amperes again, the voltage is 1.06 in the primary and 20 times that or 21 volts across the secondary. The wattage rating is $(21)^2/300$ or about 1.5 watts so that two 150-ohm 1-watt resistors are satisfactory.

The divider ratio is now 21.1/70.7 or approximately 2/7. If the losses in the core are considered, the ratio becomes almost exactly 2/7. The divider can be made up of seven 150-ohm 1-watt resistors with the tap at the 300-ohm junction as measured from ground.

If a current transformer is used in place of the series resistors, the voltage across the detectors is about 24 decibels greater and the sensitivity is increased by the same amount. The insertion loss of the circuits is not increased so that continuous monitoring is still possible. It may be necessary however to replace the crystals with tubes to prevent burnout.

The performance of the above circuits has been proved successful over as large a frequency range as 80 to 1. As a standing-wave indicator, the accuracy is within a few percent. There is sufficient power output for stability and excellent control of automatic tuning devices. Because of the low insertion loss, there is no need for the switching in and out of radio-frequency circuits. In fact, several variations for various purposes may all be left permanently attached. The chief annoyance of the circuits is the care and familiarity required in lining up a new layout where effects due to stray capacitances of as little as 1 micromicrofarad ($-j1500$ ohms at 30 megacycles) may defy all sensible attempts at correction. A helpful guide in such cases is the use of a small variable capacitance (3 to 30 micromicrofarads) that may be placed by hand across various parts while the circuit is operating to indicate whether parallel or series reactances are needed. A completed design however may be easily specified with standard tolerances so as not to require further adjustments of this sort.

Pulse Generator for the Mechanoelectronic Telephone Switching System

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IN EACH mechanoelectronic telephone switching exchange, a special group of circuits delivers the pulses that are necessary to perform the switching operations that were described in a previous article.[1]

It is obvious that these pulses must be present all the time, for any interruption or fault in pulse production will have serious and immediate consequences in the proper operation of the entire exchange. A principal objective in the development of this generator was, therefore, its reliability.

1. *Functions*

The generator must be able to deliver 5 group cycles of principal pulses designated *a*, *b*, *c*, *d*, and *e*; having the lengths and recurrence frequencies indicated in Table 1. It must be realized that group *a* is developed by 5 separate sources, each successively producing a pulse of 200-microsecond duration in a cycle that is 1000 microseconds long. Similarly, *b* has 5 sources that produce 1000-microsecond pulses to fill a 5000-microsecond cycle.

TABLE 1
GROUPS OF PRINCIPAL PULSES

Group	Number of Sources	Length of Pulses in Microseconds
a	5	200
b	5	1000
c	4	200
d	11	200
e	3	200

Due to differences among the circuits driven by these pulses, 3 types are required as shown in Table 2.

There are 2 auxiliary pulses in each 200-microsecond interval. One starts at the beginning and the other terminates at the end of each time unit. They are concerned with impulse regeneration and have the characteristics given in Table 3.

TABLE 2
CHARACTERISTICS OF PULSES

Type	Group	Base Circuit Voltage Without Pulse	Voltage Amplitude of Pulse	Top Circuit Voltage With Pulse	Rise Time in Microseconds
N	*a*, *b*, *c*, *e*	−40	+16	−24	20
N	*d*	−40	+24	−16	20
P	*a*, *b*, *c*, *e*	0	+16	+16	20
P	*d*	0	+24	+24	20
R	*a*, *b*, *c*, *d*, *e*	−110	+60	−50	10

2. *Reliability*

Several schemes may be used to ensure a reliable source of pulses.

Two generators may be provided with a suitable switching system that will replace a faulty generator with a good one.

Two disadvantages of such a system are that some time is necessary to effect the cutover, during which the pulses are either distorted or absent, and power is wasted in the standby for it must be kept under current to minimize cutover time.

TABLE 3
AUXILIARY PULSES

Type	Circuit Voltage Without Pulse	Circuit Voltage With Pulse	Pulse Position in Microseconds in 200-Microsecond Time Unit		Rise Time in Microseconds	Decay Time in Microseconds
			Start	Stop		
d_3	−48	−8	0	25	10	20
d_2	0	−24	175	200	10	20

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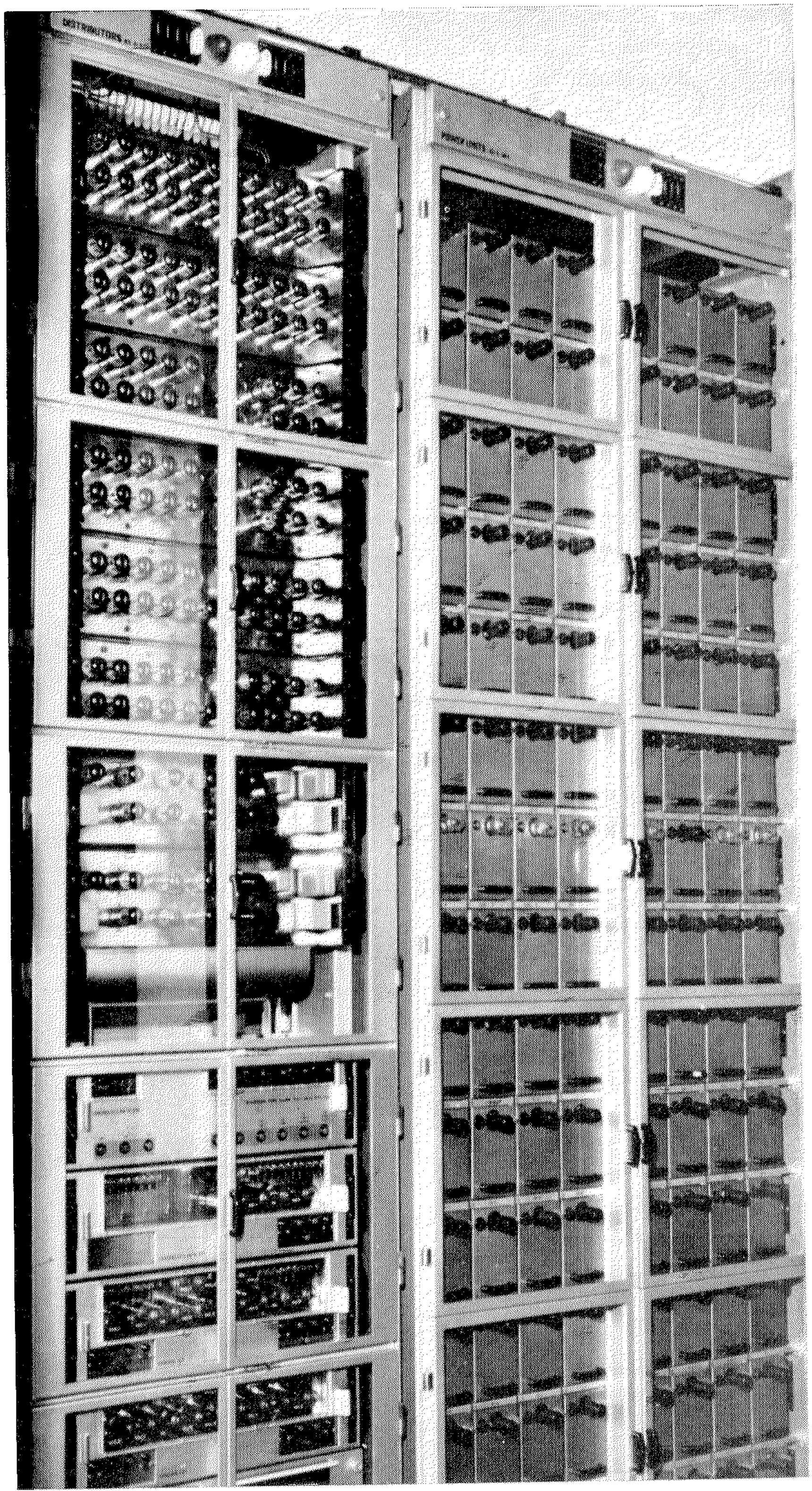
[1] J. Kruithof and M. den Hertog, "Mechanoelectronic Telephone Switching System," *Electrical Communication*, volume 31, pages 107–150; June, 1954.



These two of the three bays house the pilots, distributors, amplifiers, and control circuits.

A second scheme is to have two generators working in parallel. Their output signals go to the same points at the same time. Both generators are controlled by another circuit that is able to detect a fault in either of the two generators and remove it immediately from operation. The advantage over the previous scheme is that the switching system is protected against faulty pulses since the blocking of the improperly working circuit can be done almost instantaneously by electronic means. Furthermore, when working in parallel, the circuits share the load and no power is wasted.

Even more preferable is a scheme that would eliminate blocking. In it, the parallel working circuits are connected so that if one fails, it does not interfere with the proper functioning of the other.

In all these cases, the output of each circuit must be verified against a standard waveform to permit any discrepancy in the output signal to be detected. If it is accepted that the probability of two parallel-working circuits failing at the same time is very small, the signal of one circuit can be used to control the other, and vice versa. The probability of failure becomes smaller with simpler circuits. Therefore, if one subdivides a complete circuit into subcircuits, each of which is paralleled by a companion and associated with a fault-detecting and blocking system, the reliability of the whole system will be improved. If one of the subunits is out of order, only a small part of the complete circuit is working without parallel equipment and the chance that another fault will occur in this place is very small indeed.

Considering this fact, together with arguments of ease of construction and assembly, the pulse generator has been subdivided into 4 major parts.

A. Pilot circuit, which establishes the time base of the system.

B. Distributor or pulse-counting circuits, which produce the 5 groups of pulses mentioned in Table 1.

C. Correctors, which correct the rise and decay times of a special group of pulses that must have short rise and fall times.

D. Amplifiers, which deliver pulses with the right amount of power to the exchange.

A simplified block diagram of the generator is given in Figure 1.

All three schemes of reliability described above have been applied, their choice depending on the kind of circuits to be protected.

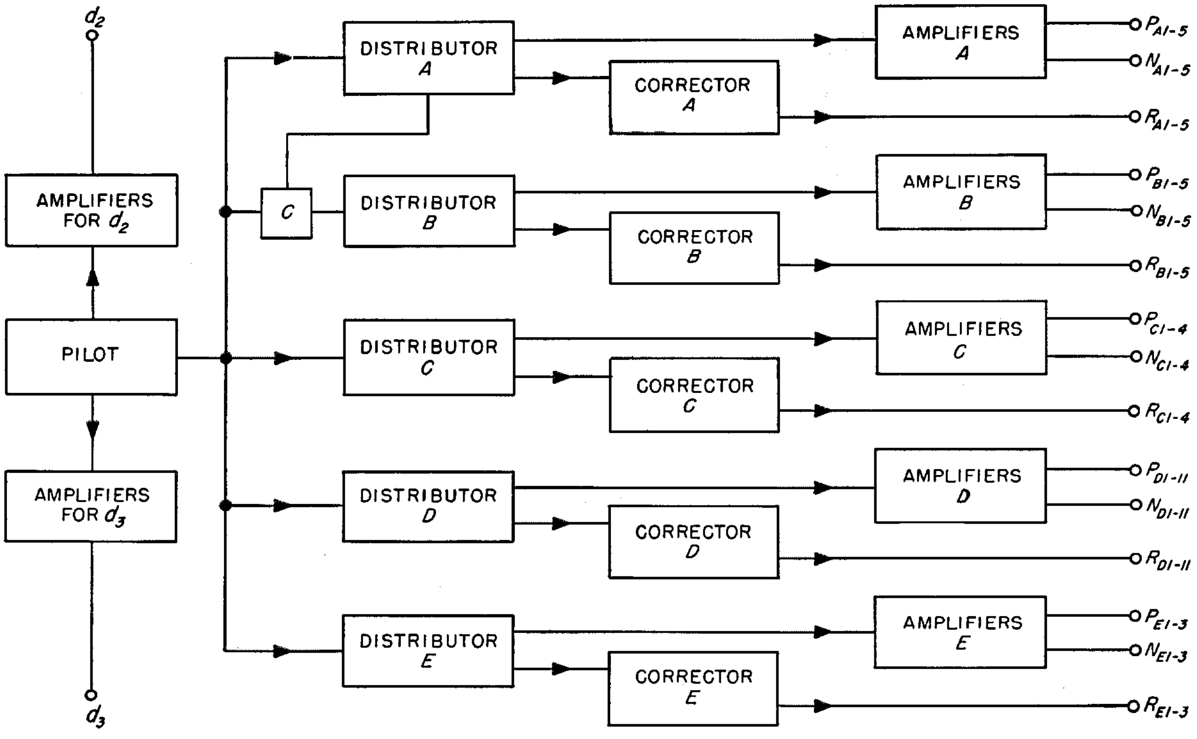


Figure 1—Block diagram of pulse generator.

3. *Description of Circuits*

3.1 Pilot Circuit

The pilot circuit has 4 functions.

A. To establish the basic recurrence frequency.

B. To furnish signals to the amplifiers for pulses d_2 and d_3.

C. To furnish pulses that are used to trigger the distributor circuits.

D. To furnish pulses that monitor the operation of the pilot circuits.

To ensure reliability of operation, two pilots are operated in parallel with a common control that electronically blocks either circuit if it does not deliver a proper signal.

The circuits make use of conventional multivibrators, flip-flops, amplifiers, et cetera so that an explanation of their functions will be based on block diagrams only.

Referring to Figure 2, each pilot is composed of the following parts.

A. Generator Gd_2 delivers pulses of 25-microsecond duration at intervals of 200 microseconds. This is a self-starting multivibrator controlled by a loosely coupled resonant circuit that is adjusted to produce a signal of 5000 cycles per second.

B. Separating or isolating tube S_1 that delivers

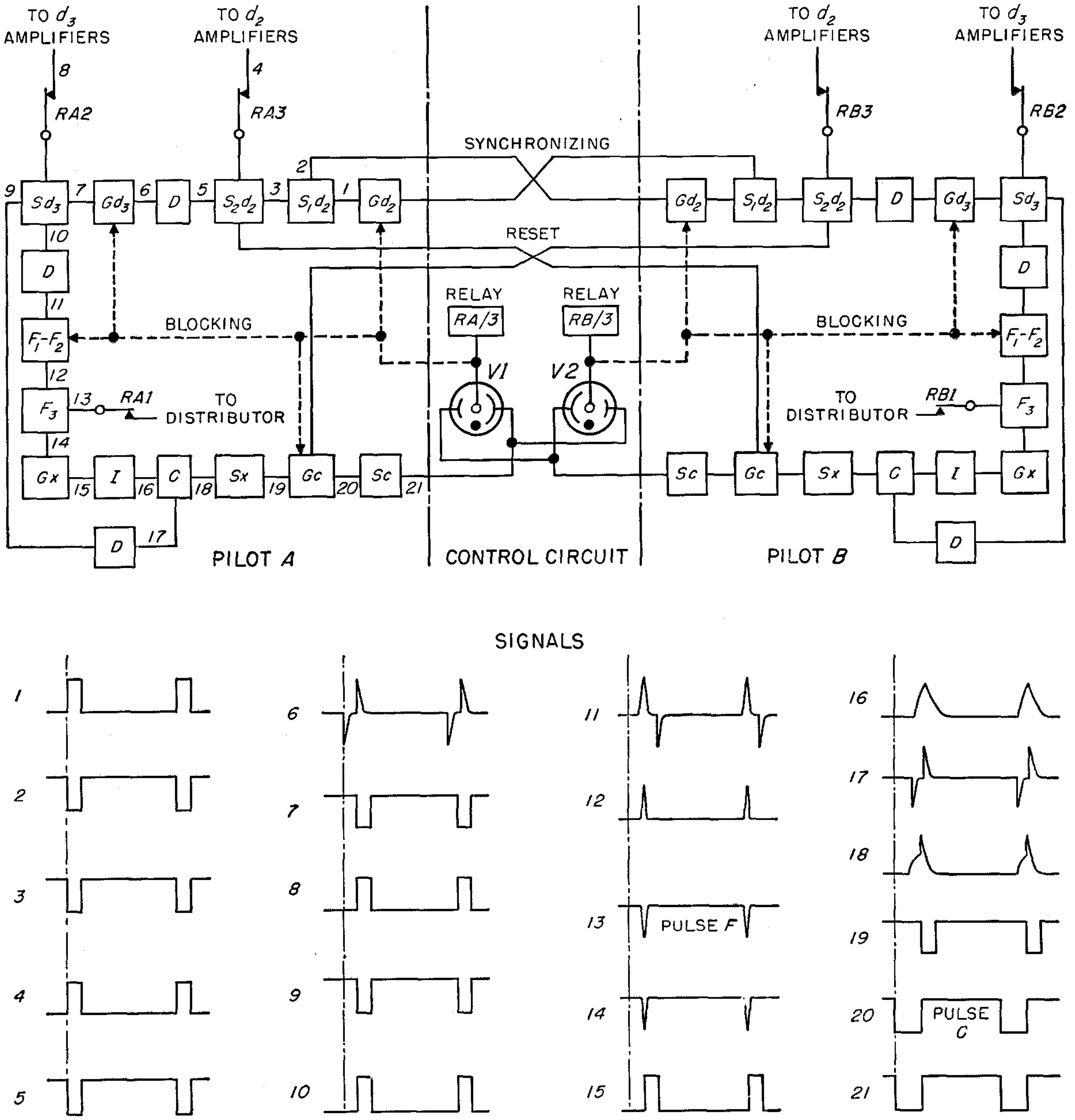


Figure 2—Pilot and pilot-control circuits. C = combination, D = differentiation, $F-F$ = flip-flop pulse generator, G = generation of indicated series of pulses, I = integration, and S = separation or isolation circuit.

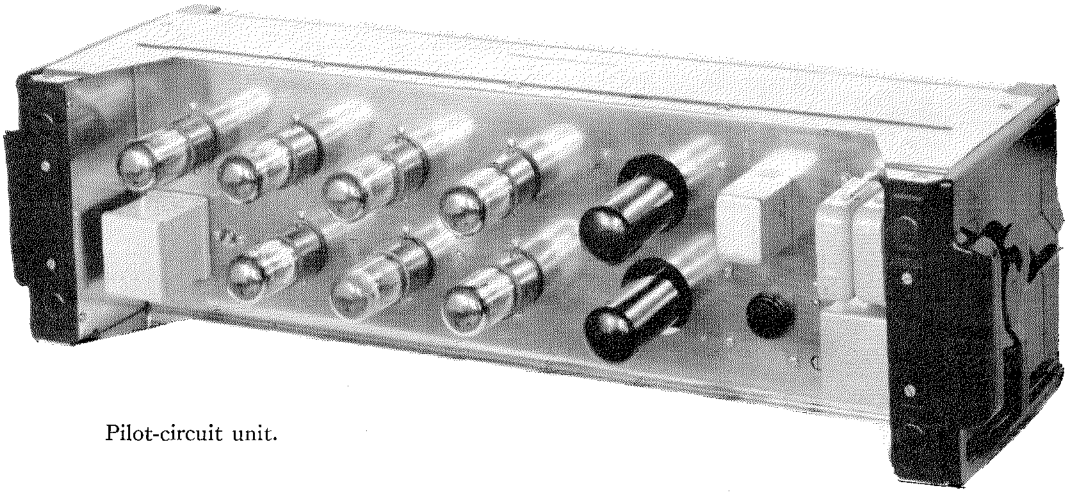
negative synchronizing pulses to the other parallel-working pilot.

C. Separating tube S_2 delivering positive pulses to the d_2 amplifiers and a reset signal to the parallel-working pilot for control purposes.

C. Generator Gd_3 is triggered by the differentiated signal coming from S_2 and D; it sends a negative pulse to a separating tube.

E. Separator Sd_3 sends pulses to the d_3 amplifiers.

F. Generator F_1-F_2 is triggered by a differentiated signal coming from Sd_3 and sends a short signal to the distributor circuits through amplifier F_3.



Pilot-circuit unit.

G. Generator Gx is triggered by F_3 and furnishes pulses that are integrated and combined with differentiated pulses coming from Sd_3.

H. To open separator tube Sx, it is necessary that both the integrated and the differentiated signals be present at its input and that a proper time relation exist between these signals. This means that if the trailing edge of d_3 comes too early or too late, due to malfunctioning of Sd_3, the differentiated signal does not ride high enough on top of the integrated pulse and the combined signal is unable to open tube Sx.

I. Generator Gc is a monostable multivibrator with a restoring time greater than 150 microseconds. However, it is forced to restore after this time lapse by a signal coming from the S_2d_2 tube of the other pilot and is triggered by the signal coming from tube Sx of its own circuit. The reason for applying forced restoration in Gc will become clear after the description of the control circuit.

J. Tube Sc serves as a separator tube between generator Gc and the control circuit.

This control circuit consists mainly of two cold-cathode tubes, each associated with a relay that is operated when the corresponding tube is fired. The function of the relay is to open all wires connecting a faulty pilot to the remaining part of the equipment so that it may be removed without disturbance and to give an alarm.

Two pulses are applied to each cold-cathode tube. The C pulse coming from pilot A is applied to the cathode of tube $V1$ and to the trigger of tube $V2$, while the C pulse coming from pilot B is applied to the cathode of tube $V2$ and to the trigger of tube $V1$.

The result is that if both pilots are working under normal conditions, the potential difference between cathode and trigger remains insufficient to fire either of the tubes, although both electrodes are regularly making potential excursions. Only if one of the control pulses is absent will the conditions in the corresponding cold-cathode tube be favorable to firing. This also explains why forced restoration has been applied on Gc.

If, for instance, this tube were a one-shot multivibrator, it would be possible due to small changes in its components for the control pulse delivered by it to be shorter or longer than that delivered by the other pilot. If this difference becomes large enough, one of the cold-cathode tubes would be actuated. It could be the one corresponding either to the faulty pilot or the good pilot, since there is no telling which control pulse is too long or too short.

It is interesting to examine some of the major faults that could occur.

A. *Generator* $F_1 - F_2$ *stops working.* Signal *12* is not produced and thus signal *16* is not available to combiner *C*. Since *Sx* cannot be opened by signal *17* alone, the cold-cathode tube in the corresponding control is operated.

B. *Length of pulse* d_3 *is shortened or lengthened.* Signal *17* comes too early or too late with respect to signal *16* and the combined signal *18* is not large enough to open tube *Sx*, with the same result as above.

C. *Generator* Gd_3 *stops working.* Signal *7* is absent and again the corresponding cold-cathode tube is fired.

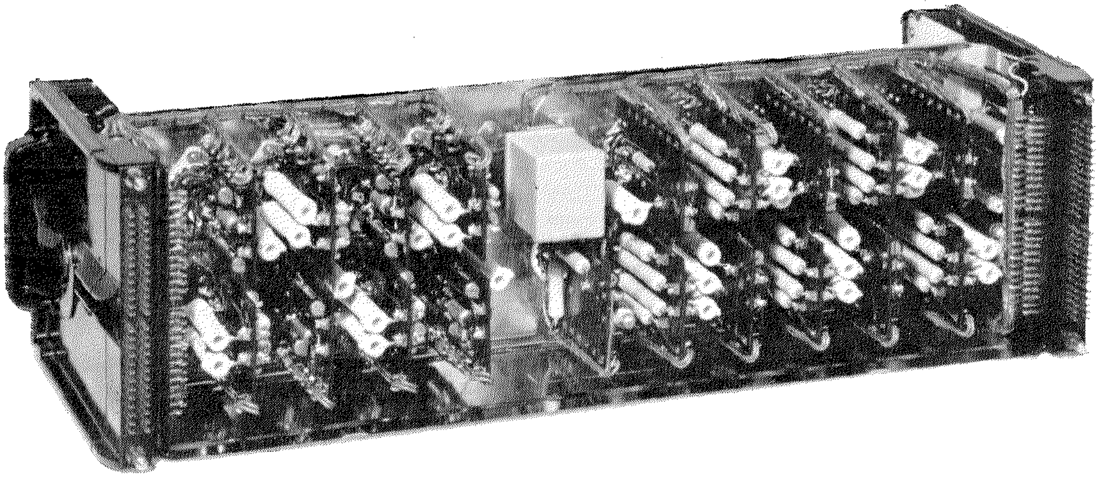
D. *Generator* Gd_2 *is stopped.* None of the pulses is present and again the control is actuated.

In each case, the cold-cathode tube is fired in the same time unit that the failure occurs. Blocking is exerted immediately on all crucial points of the defective pilot, so that no faulty pulses may be sent to the following circuits.

The electronic blocking system makes use of a point in the control circuit where an important voltage drop is observed when the control is operated. This low voltage is applied through rectifiers on the grid circuit of the most important tubes in the pilot, so that these tubes are unable to emit signals.

A system of push-buttons permits the intentional firing of a control, as well as the extinguishing of the tube after a faulty unit has been repaired and replaced.

The circuit arrangement is such that the tubes cannot deionize and the operated relays cannot be released unless the circuit is restored to working order.



Rear view of a *b*-group distributor. The plug-in contacts may be seen at each end.

3.2 Distributor Circuits

The purpose of the distributor circuits is to generate the 5 groups of pulses shown in Table 1. Each distributor has two identical units working in parallel under the supervision of a common control.

Each of the two units receives trigger pulses from both pilots, so that even if one pilot is out of order all of the distributor units continue to function.

A distributor is a circuit of the well-known ring counting chain type having as many flip-flops in a closed loop as there are sources in a cycle, for example, 5 for cycle *a*, 11 for cycle *d*, et cetera.

The four units delivering 200-microsecond pulses for cycles *a*, *c*, *d*, and *e*, are directly triggered by pulses coming from the pilot circuit through amplifier F_3 and have the waveform of signal *13* in Figure 2. Since these trigger pulses are produced every 200 microseconds, the output pulses are automatically of the correct length as each output pulse is derived from one of the flip-flop circuits during the time between adjacent trigger pulses. The counting chains are arranged to have the required stepping cycle to provide the number of sources required.

Cycle *b*, which has to deliver pulses of 1000-microsecond duration, is triggered by the same signal as the other chains. After the 1st trigger pulse has passed, a gating circuit controlled by an *a*-cycle pulse prevents the 4 following trigger pulses from actuating the flip-flop, which holds until the 6th pulse of the *a* cycle arrives. Since an *a*-cycle pulse is present only once every 5 time units, a trigger pulse actuates the *b*-cycle units only once every 1000 microseconds.

Each flip-flop is followed by two separator tubes, which are cathode-followers. The purpose

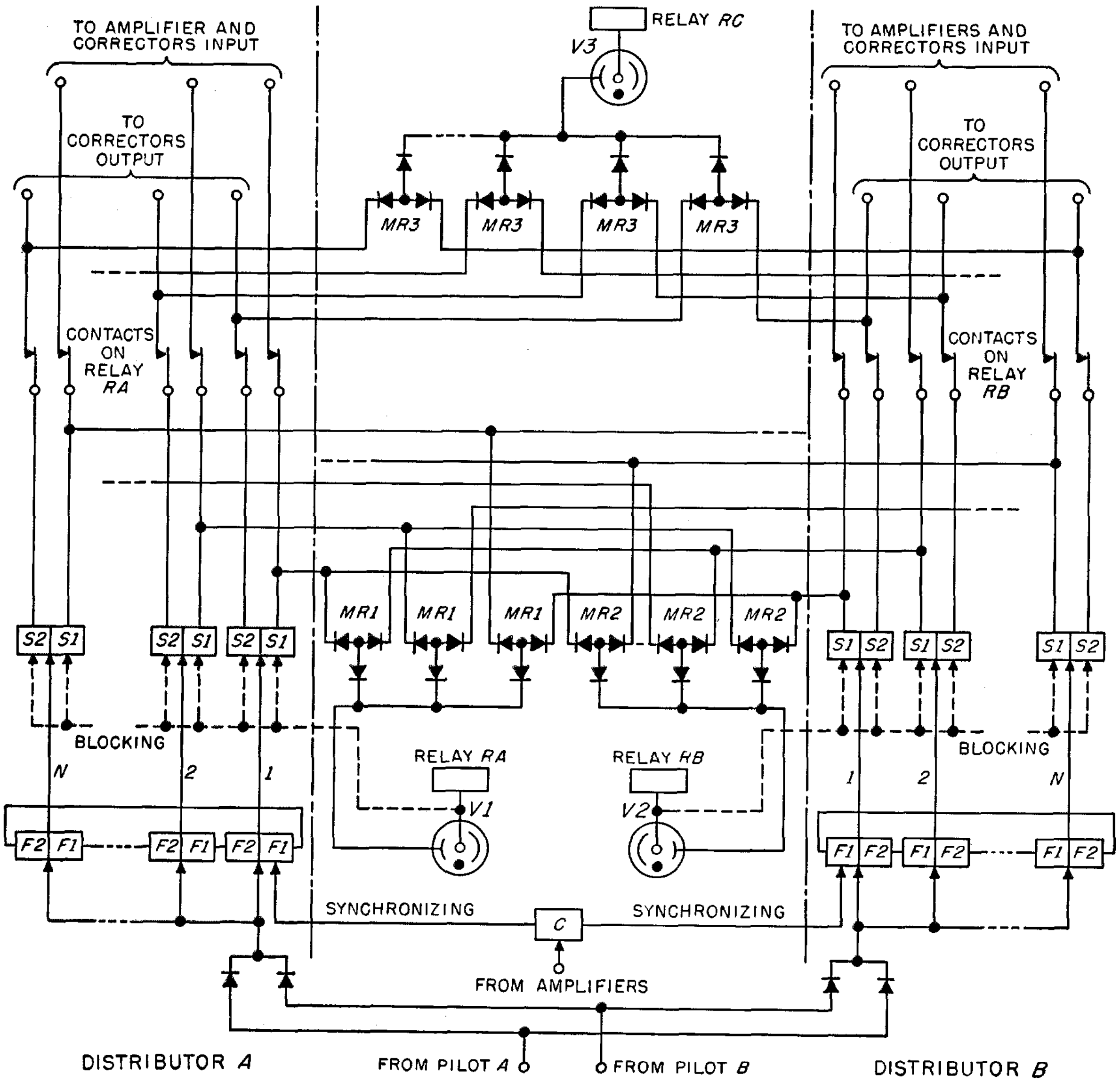


Figure 3—Two distributors are operated in parallel under the supervision of a common control circuit.

of these tubes is principally to act as buffers between the counting chains and the other circuits of the generator.

In Figure 3, two parallel-working distributor chains are represented with their control circuit. Two kinds of controls are introduced into the distributors; the first is applied to the counting chains proper and the second to the separator tubes.

A major fault that could occur in a distributor would be for the counting chain to stop stepping around. In that case, one of the flip-flops will be in the actuated state while all the others are in the nonactuated condition. This effect is inherent in the design of the counting chains. If, therefore, a chain does not continue to step, it may be accepted that one of the outlets will always be at a much higher potential than all others.

For control purposes, each pulse circuit of distributor *A* has been associated with the pulse circuit of following time rank produced by distributor *B*, and vice versa. The circuits are isolated from each other by the networks of rectifiers *MR1* and *MR2* that are connected to the triggers of the cold-cathode tubes *V1* and *V2* that can disable the distributors.

When the two chains are working under normal conditions, no signal appears on the triggers of the control tubes since one side of each rectifier group *MR1* and *MR2* is always at a low potential. For instance, when *S1* on point *1* of distributor *A* is high, the other side of *MR1* is low, since no pulse is delivered on point *2* of distributor *B* at that instant. It is only if one of the flip-flops in a chain does not turn over, that both sides of a rectifier group *MR1* or *MR2* will be high at the same time and the corresponding cold-cathode tube will be fired.

If, for instance, distributor *A* stops working when its outlet *1* is high, the control tube of this distributor will be triggered in the following time unit when the pulse from distributor *B* appears at the other branch of the rectifier combination *MR1*. This operates *V1* and disables the faulty distributor *A*. If it is distributor *B* that stops on one of its points, a pulse of following time rank from distributor *A* appearing on the second branch of *MR2* will cause the proper cold-cathode tube to light and disable distributor *B*.

Blocking is exerted on all separators of the faulty chain, so that no signal will be emitted by this chain. At the same time, a relay is operated to open all outlets and give an alarm. Blocking is done by lowering the potential of the grids of the separators, so that they are brought below cutoff.

The control of the separator stages has been accomplished by grouping by pairs the outputs of tubes *S2* of the same rank and by strapping these groupings through separating rectifiers *MR3* to the trigger of the cold-cathode tube *V3*. In this way, the potential observed at this trigger will be equal to the top of the pulses since both sides of each grouping are always high at the same time. This is repeated for each grouping in turn so that there appears on the common point a signal that is equal to all pulses placed side by side. If, however, a separator fails to work, one side of a group of *MR3* will be low and a gap is produced in the signal on the common point. Tube *V3* is fired and relay *RC* is operated, giving an alarm. No blocking is exerted since a failing separator cannot influence the parallel-working circuits, all of which are equipped with separating rectifiers.

To keep both counting chains in step, a synchronizing signal is injected on the first flip-flop of each chain. This signal is obtained from the corresponding amplifier stages after proper transformation in circuit *C*.

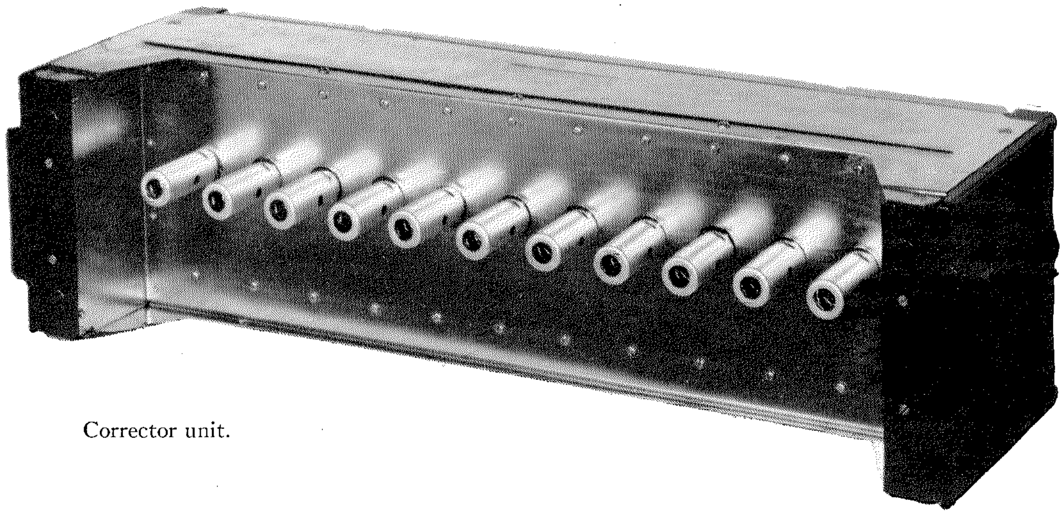
Here again a push-button system permits the intentional blocking and unblocking of a distributor.

3.3 CORRECTOR CIRCUITS

The purpose of these circuits is to produce the *R* type of pulses for which the requirements regarding rise time and decay are much more exacting than for the other pulses. The power with which these pulses should be delivered is negligible so that they are taken directly from the second separator tubes in the distributor circuits and sent to the different points in the exchange where they are required. This means long wires, which present the problem of distributed capacitance and the correction of the waveform distortion that it produces.

The principle on which the correction of these pulses is based consists in the forced charge of the wire capacitance at the beginning and the forced discharge of this wire at the end of a pulse. To effect these charges and discharges, use is made of thyratrons that serve as ultrafast switches while capacitors are used to store the charges that have to be injected on the wires.

Figure 4 shows a schematic drawing of a 3-point corrector, without its control circuit.



Corrector unit.

The distributed capacitances of the wires connected to the pulse leads are represented by *C1*, *C2*, and *C3*, while the storage capacitors are *C1′*, *C2′*, and *C3′*.

The pulses coming simultaneously from distributors *A* and *B* are connected to the 3 *R*-pulse leads via separating rectifiers *S1* and *S1′*.

The storage capacitors are connected between the pulse leads through thyratrons that are triggered by pulses coming from the distributors after they have passed through a pulse-forming circuit. The thyratron is thus triggered by a signal coming from the distributor at the moment another pulse from the distributor appears on the corresponding pulse lead.

Let us suppose that a pulse is present on output *VS2–1*. This lead is, therefore, at about −50 volts since this is the top value of the *R* pulses. Rectifier *S2* is blocked on that point while all other leads are at −110 volts since their rectifiers *S2* are passing current.

When the pulse on lead *1* stops, the next pulse starts on lead *2* and at the same time *V2* is triggered. With the thyratron conducting, a discharge path is established from the wire capacitance *C1* through lead *1*, storage capacitor *C1′*, resistor *R2*, tube *V2*, lead *2*, and wire capacitance *C2* to ground. Resistor *R2* limits this discharge to a value suitable for the tube. The result is that the capacitance of wire *1* is discharged while that of wire *2* is now fully charged. The operation repeats at the end of the pulse on wire *2* in favor of wire *3*, and continues around the chain.

The scheme used to insure reliable operation in the pilot and distributor circuits is not applicable here. The line capacitance is really a part of the circuit and this makes it impracticable to apply a parallel-working principle.

There are, therefore, two corrector units for each cycle, one of which works into the live load while the other is connected to a dummy load.

If the first circuit develops a fault, the live load is switched in place of the dummy load at once, while if the standby circuit misfunctions, only an alarm is given to call attention to the trouble.

The control circuit comprises a fault-detecting system consisting of rectifiers and cold-cathode tubes, and controls both the top and the bottom of the pulse train coming out of each circuit. Again push buttons are provided for intentional firing and deionizing of the control tubes.

3.4 POWER AMPLIFIERS

The *P* and *N* pulses must deliver a certain amount of power to the exchange. The entire equipment is designed to produce more power for each outlet than necessary so that in case of failure of some apparatus the power delivered will not be less than the minimum amount required to drive the exchange.

To satisfy this requirement, an amplifier was constructed that could deliver a unit amount of power. This amplifier was kept as simple as possible, comprising only one tube, an output transformer, and some associated small components.

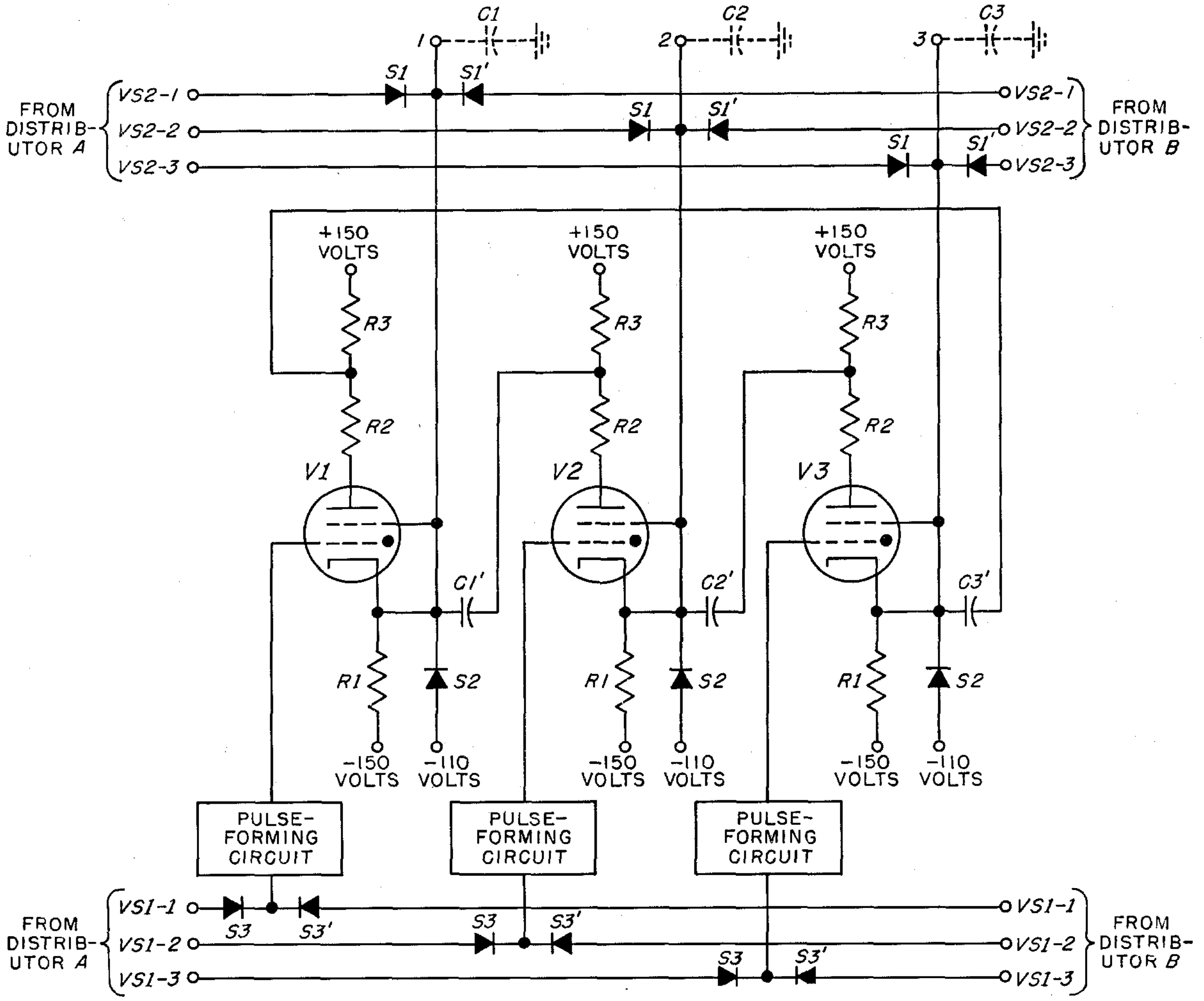


Figure 4—Three-point corrector.

On each output lead are placed as many of these unit amplifiers in parallel as are required for the amount of power to be delivered, plus one additional amplifier. In case of failure of one of the amplifiers, there will still be enough in working condition to carry the load.

As may be seen from Figure 5, the *P* and *N* pulses are obtained at the output of the same amplifiers. The output transformers have two secondary windings to permit the base of the pulses to be placed at the correct value; that is, ground for *P* pulses and −40 volts for *N* pulses. Resistor *R3* permits adjustment of the height of the pulses.

The control circuit for an ampilfier comprises rectifiers *S1* and *S2* and resistors *R1* and *R2*. When no pulses are delivered, the output lead for *N* pulses is at −40 volts, while the output lead for *P* pulses is at ground potential. These potentials are impressed on control wires *C1* and *C2* through rectifiers *S1* and *S2*, which are conducting.

When a pulse arrives, both control wires are raised by 16 volts as long as the pulse is present, returning to their original values at the end of the pulse. In other words, the control pulses are a reproduction of the *P* and *N* pulses. All pulses from a cycle are brought to a common control circuit, by means of separating rectifiers *S5* and *S6*, so that the input of the control transformer *TC* sees a constant voltage that is equal to the top value of the pulses.

If, however, one of the amplifiers does not deliver signals on one of its two outputs, *N* or *P*, the corresponding control wires *C1* or *C2* are kept at −40 volts or ground by means of rectifiers *S1* or *S2*, so that the control pulses that are normally delivered by the parallel-working amplifiers do not appear on the control wires. The result is that one pulse of the train applied to transformer *TC* is missing, and tube *VC* fires. An alarm is given but no blocking is exerted since a defective unit is separated from the others by means of rectifiers *S3* and *S4*.

Here again each amplifier receives input pulses from two distributors so that even with one of these circuits out of order, pulses continue to be delivered normally to the exchange.

4. *Construction of Pulse Generator*

4.1 Components

Reliability again has been the major factor in choosing the components of the pulse generators.

All resistors and capacitors have been selected to have large safety factors regarding power dissipation and working voltages. Possible faulty conditions have been considered in determining these values, so that failure of one component will not subject other parts to excessive loads and require their replacement also.

The number of germanium rectifiers has been

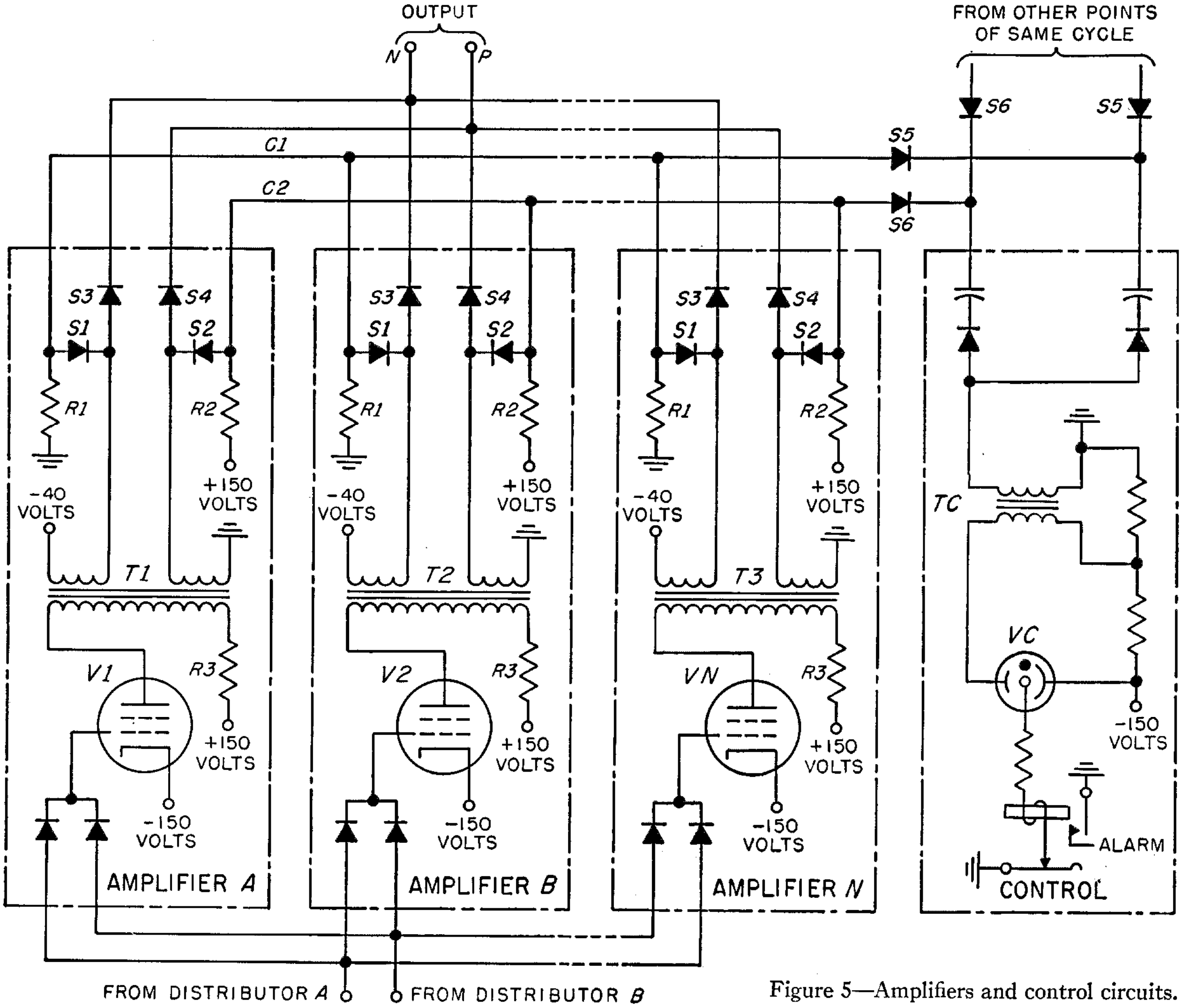


Figure 5—Amplifiers and control circuits.

kept to a minimum because of the rather high price of this component. Most of the rectifiers are of the miniature selenium type and generous safety factors have been applied regarding the voltage each disk must withstand.

Four different types of tubes are used in the generator. The *13D1* Brimar tube manufactured by Standard Telephones and Cables was selected as the twin triode principally because with these tubes the contacts of the jack-in units do not have to carry heavy currents, since a 25-volt filament supply is required.

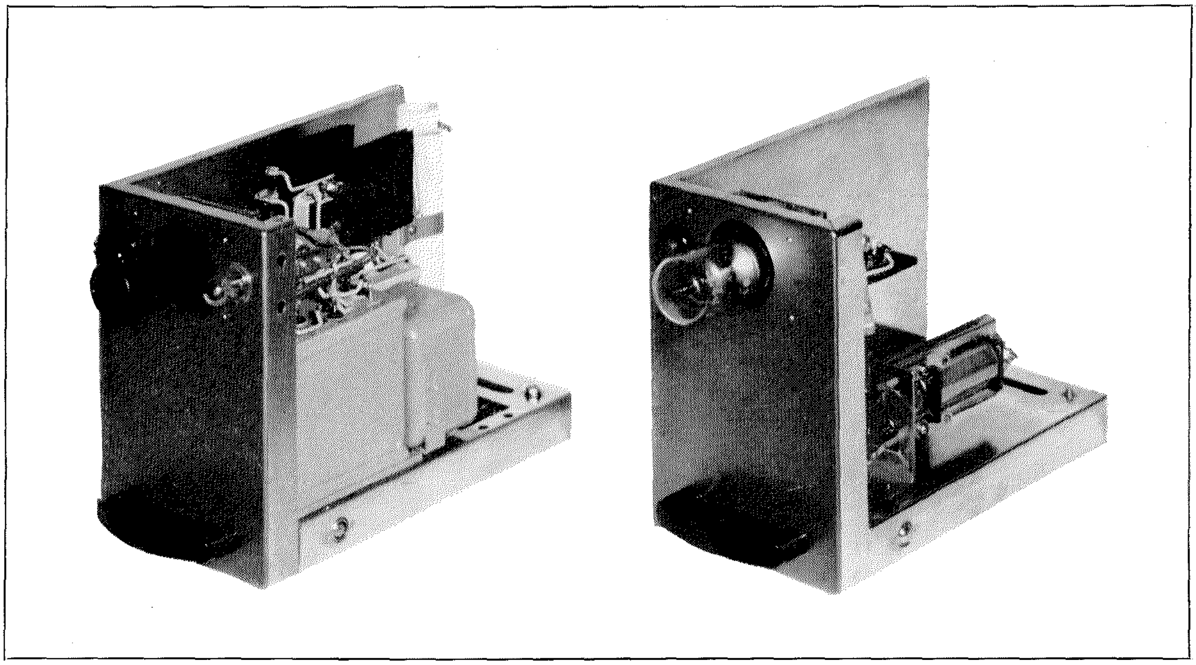
Ordinary *6V6* tubes were chosen for the power stages, while in the pilot circuit, a few *6L6*'s are used; both of these tubes are highly reliable when applied with proper care.

The thyratrons in the corrector circuits are of the *2D21* type. Although of miniature size, they are able to pass the required peak currents. All tubes are subjected to a short life test to eliminate early failures.

4.2 General Layout

The complete pulse generator comprises three bays of equipment. Two of the bays are of the type used in the *7E* system and employ the same plug-in units that are noted for having very reliable high-pressure contacts.

In each individual unit, all the tubes are placed on the front side with their sockets mounted on a plate that separates the tube envelopes from the components in the back; in this way heat transfer to these components is kept to a minimum.



Amplifier unit and its control system.

The amplifier bay houses about 100 unit amplifiers. It is constructed along the same lines as the switch bays used elsewhere in the exchange.

The bays are enclosed by doors and air is forced to flow from the bottom to the top. This is necessary to keep the ambient temperature in which the components are working at a normal value. Cooling the vacuum tubes considerably lengthens their useful life.

The air intake is outside of the exchange building. A filter is included in the airflow system so that all dust particles are eliminated.

4.3 Power Supply

The voltage sources to the generator are under the control of timing circuits for the purpose of applying these voltages in the correct order.

To reduce the danger of cathode damage, high voltages are only connected when the filaments of the tubes are warm. The control circuits are the last to be energized so that they are unable to work during the starting interval of the generator.

An interruption of the primary power supply causes all high potentials to be removed at once. They are again applied in the proper sequence after the supply is re-established. The time lapse for the applications of the different voltages is directly related to the duration of the interruption. If it is a very short one, the tubes have had practically no time to cool off and the high potentials are reconnected very rapidly whereas with longer interruptions the timings are automatically lengthened.

The filament transformers are so connected that even with one of them out of order, there is still at least one of each type of unit, namely: pilot, distributor, and corrector, that continues to work normally. As for the amplifiers, the filaments are so distributed over the transformers that even with one out of order there are enough amplifiers left on each point to supply the exchange.

5. Conclusion

The pulse generator is a major part in the mechanoelectronic exchange and the slightest fault in pulse delivery will have immediate consequences in the functioning of the exchange.

It is, therefore, understandable that some time and effort were devoted to the engineering and construction of this circuit, with absolute reliability of operation as a primary objective. It is felt that this goal has been achieved.

Recent Telecommunication Development

Etching of E. H. Armstrong

Edwin Howard Armstrong (1890–1954) is portrayed in the latest of the series of etchings published by the International Telecommunications Union. Born in the city of New York, he was educated at Columbia University with which he was closely associated until his death.

Major Armstrong contributed four outstanding inventions. His regenerative circuit of 1913 supplied a much-needed increase in sensitivity and selectivity to the early vacuum-tube detector. In 1918, he produced the superheterodyne that is the basis of practically all modern radio receivers. It allows unlimited increases in selectivity and sensitivity without impracticable multiplicity of controls or instability. It makes the noises developed in the circuits preceding the amplifier the real limitation on over-all gain. Two years later, superregeneration promised to supplant all other methods of reception, particularly at the higher frequences, but some inherent limitations restricted its applications severely. In 1933, his method of wide-band frequency modulation provided the first practical system of eliminating the effects of static; it makes the acoustical output of thunder more damaging to reception of music during a local thunderstorm than the electrical disturbances of the lightning that produces the thunder.

The etching of Armstrong is the twentieth in the series that was started in 1935. On a good grade of paper measuring 9 by 6$\frac{5}{8}$ inches (23 by 17 centimeters) including margins, these etchings are available at 3 Swiss francs each from Secrétariat général de l'Union internationale des Télécommunications, Palais Wilson, 52, rue des Pâquis, Genève, Suisse. The entire series is comprised of etchings of Ampere, Armstrong, Baudot, Bell, Erlang, Faraday, Ferrié, Gauss and Weber, Heaviside, Hertz, Hughes, Kelvin, Lorentz, Marconi, Maxwell, Morse, Popov, Pupin, Siemens, and Tesla.

United States Patents Issued to International Telephone and Telegraph System November, 1954—January, 1955

United States patents numbering 18 were issued between November 1, 1954 and January 31, 1955 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that are of more-than-usual interest are included.

W. A. Billings and Z. Brzozowski, Joining Wire, 2 696 227

A. Brown and D. A. Romlin, Electric Switch, 2 696 553

G. Deakin, Telephone System, 2 696 525

M. den Hertog and C. de Zeeuw, Selection Systems for Electrical Circuits or Equipment, 2 694 751

M. den Hertog, System of Identification of the Calling Line in Telephone Systems, 2 694 753

M. J. DiToro, W. Graham, and B. M. Dwork, Fundamental-Pitch Detector System, 2 699 464

J. H. Eisele and A. R. Geisselsoder, Dry-Disk Rectifier Assembled from Unperforated Rectifier Plates, 2 698 918

R. Haberkorn, Pneumatic-Tube System, 2 698 722

P. L. Hancock, Sideband Generator, 2 697 220

J. V. Martens, Voice-Frequency Receiver, 2 698 878

S. E. Mayer and H. F. Sterling, Electric-Current Rectifier, 2 695 380

H. G. Miller and J. Stillman, Automatic Tuning Device, 2 699 504

W. Reinhard and A. Rappold, Anode-Voltage Supply for Television Receivers, 2 694 784

W. Sichak and H. A. Augenblick, Jr., Tunable Band-Pass Filter, 2 697 209

F. H. Stieltjes, Balancing of Carrier Cables, 2 696 526

V. J. Terry, R. H. Dunn, and P. G. Collier, Remote Control and Supervision System, 2 694 802

E. P. G. Wright, D. S. Ridler, D. A. Weir, and J. Rice, Telecommunication Exchange, 2 694 752

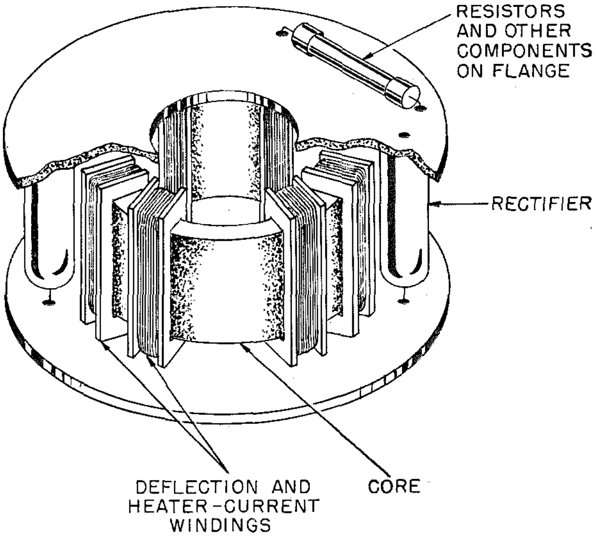
E. S. Wright and A. C. Robertson, Piezoelectric Crystal Plate, 2 698 909

Anode-Voltage Supply for Television Receivers

W. Reinhard and A. Rappold

2 694 784—November 16, 1954

A deflection yoke for television tubes that utilizes a toroidal magnetic core with the deflection windings magnetically coupled to auxiliary coils also wound on the core for supplying the heating current for the rectifiers used to obtain the cathode-ray-tube high voltage from the deflection waveforms. These rectifier tubes are mounted in the same yoke structure and since the heating current is obtained at this high-voltage point, it is not necessary to provide special insulation for them.

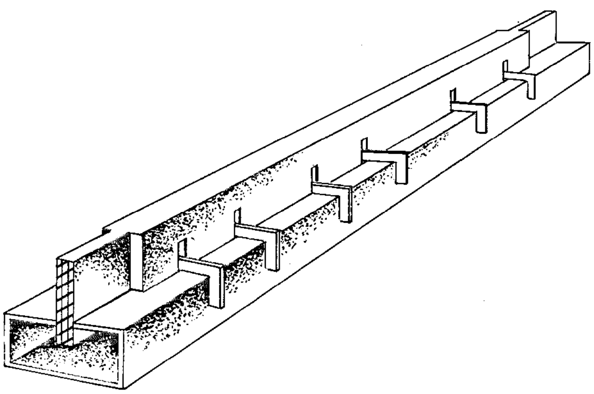

RESISTORS AND OTHER COMPONENTS ON FLANGE
RECTIFIER
DEFLECTION AND HEATER-CURRENT WINDINGS
CORE

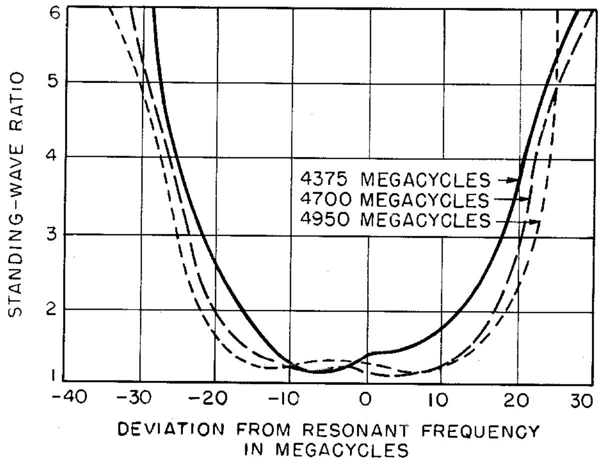
Tunable Band-Pass Filter

W. Sichak and H. A. Augenblick, Jr.

2 697 209—December 14, 1954

A tunable waveguide-type microwave filter in which the filter is divided by irises into sections and all of these sections are tuned simultaneously by the insertion of an insulating strip into the waveguide along its entire length.





Balancing of Carrier Cables

F. H. Stieltjes

2 696 526—December 7, 1954

A method of balancing high-frequency cables to compensate for phase disturbance causing crosstalk in which the conductors of successive sections of the cable are systematically exchanged in groups, pairs of groups being transposed with each other on adjacent sections of the cable.

Electric-Current Rectifier

S. E. Mayer and H. F. Sterling

2 695 380—November 23, 1954

A rectifier formed of titanium dioxide of the conducting type having on its upper surface a thin layer of nonconducting titanium dioxide. The base-plate electrode is silver on the surface of the conductive titanium dioxide and the counterelectrode deposited on the nonconductive titanium dioxide may consist of lead foil, tellurium, or may incorporate thallic oxide.

Pneumatic-Tube System

R. Haberkorn

2 698 722—January 4, 1955

A deflection switch is described for pneumatic-tube carriers in which there is provided an arrangement for simultaneously shunting the carrier into a desired branch tube and also controlling the flow of air by means of a damper.

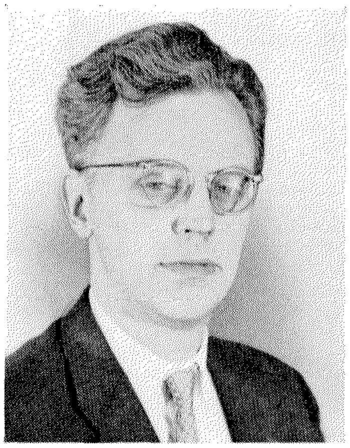
Fundamental-Pitch Detector System

M. J. DiToro, W. Graham, and B. M. Dwork

2 699 464—January 11, 1955

A system for producing pulses repeated in accordance with the fundamental frequency of various voice sounds. The voice sounds are rectified; amplitude discrimination is incorporated and band-pass filters determine the fundamental frequency of the sound. This is accomplished by various switching and storage systems and is followed by a differentiating circuit for producing pulses timed in accordance with the fundamental frequency.

Contributors to This Issue



Robert T. Adams

Robert T. Adams was born in Sparkill, New York, on January 6, 1915. He received his engineering education at Cornell University.

From 1935 to 1937, he did engineering work for the National Rayon Dyeing Company. For the next eight years, he was with the Western Electric Company.

Since 1946, Mr. Adams has been associated with Federal Telecommunication Laboratories and is now a department head. His work on air navigation, communication receivers, microwave links, and radar has resulted in several patents. He is coauthor of a paper in this issue on reflectometers.

Mr. Adams is an Associate Member of the Institute of Radio Engineers.

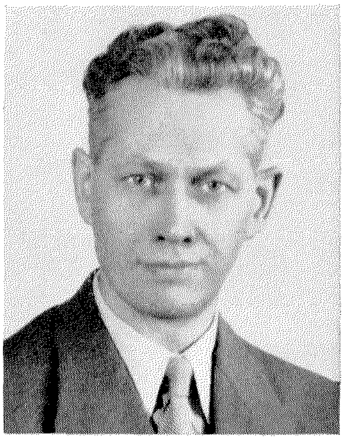
• • •

Reginald B. W. Cooke

Reginald B. W. Cooke was born in London, England, on May 17, 1913. He joined Standard Telephones and Cables, Limited, in 1929 and until 1943 was a production-planning engineer.

After service with the Royal Corps of Signals, he joined Standard Telecommunication Laboratories in 1946 and worked with A. H. Reeves on a number of projects that included gas counter valves and transistor and other semi-conductor devices. He is co-author of the paper in this issue on a positive-gap diode.

• • •



J. L. Donceel

J. L. Donceel was born in Antwerp, Belgium, on August 12, 1899.

He joined the apparatus division of Bell Telephone Manufacturing Company in 1920.

He was engaged successively in the inspection, investigation, and development of automatic switches.

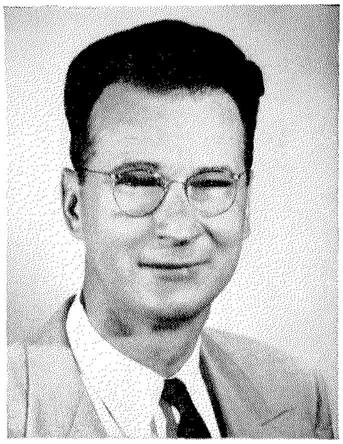
In 1938, Mr. Donceel was placed in charge of the life-test department and in 1945 of the design of automatic apparatus. He was appointed head of the apparatus division in 1953. He reports here on a new step-by-step selector switch.

• • •

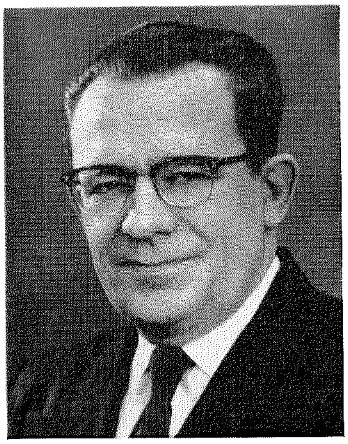
Alexander Horvath was born in New York City on December 2, 1911. He received a bachelor's degree in physics from Antioch College, where he participated in the spectroscopic analysis of the chlorophylls and their derivatives.

After a short period of instructing at Antioch and later as an instructor and field engineer with the United States Army Signal Corps, he joined Federal Telecommunication Laboratories in 1944. He is now in charge of a components department of the guided-missile laboratory and is a coauthor of a paper on reflectometers in this issue.

Mr. Horvath is a Senior Member of the Institute of Radio Engineers.



Alexander Horvath



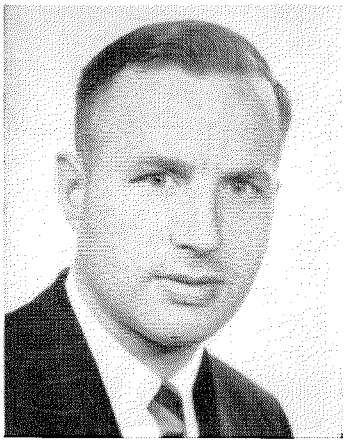
Joseph F. Houdek, Jr.

Joseph F. Houdek, Jr. was born on May 10, 1908 in Chicago, Illinois. He received a BS degree in physics and chemistry from Lewis Institute in 1934.

From 1934 to 1937 as a consulting engineer and orchestra leader, he was concerned with crystal microphones as orchestral pickup devices and the design of associated amplifiers. From 1937 to 1941, he was with the General Electric X-ray Corporation.

In 1941, Mr. Houdek joined the laboratories of Kellogg Switchboard and Supply Company, becoming chief of the general laboratory in 1954. In this issue, he describes the design and performance of capacitor microphones.

Mr. Houdek is a registered professional engineer in Illinois, a member of the American Institute of Electrical Engineers and active in its Chicago Section, a member and officer of the Chicago Acoustical and Audio Group, and holds memberships in the Armed Forces Communication and Electronics Association and in the Physics Club of Chicago. He has been granted several patents on wire-communication apparatus. In addition to his engineering work, he has been an active musician for over 30 years.



John W. Howes

• • •

John Wesley Howes was born at Neutral Bay, Sydney, Australia in 1916. He attended Sydney Technical College.

After doing radio receiver service work he joined the Australian army in 1940, transferring to the Royal Australian Air Force and attaining the rank of Flight Lieutenant.

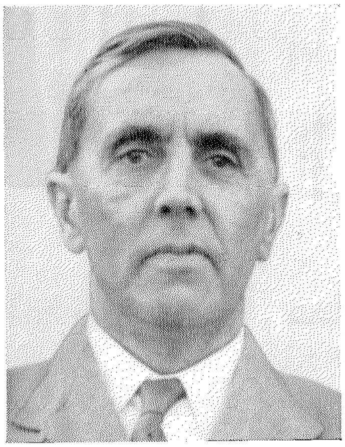
Late in 1945, he became a member of the engineering staff of the valve laboratory of Standard Telephones and Cables Pty., Limited, where he is concerned with the manufacture of transmitting valves. He reports in this issue on thermistors.

Mr. Howes is a member of the Institution of Radio Engineers, Australia.

• • •

Alec Hurley Reeves was born in Redhill, Surrey, England, in 1902. He received the DIC from Imperial College of London University and is also an associate of the City and Guilds Institute. He is a member of the Institution of Electrical Engineers.

In 1923, he joined the International Western Electric Company, which later became Standard Telephones and Cables, Limited. He worked on the original transatlantic radiotelephone system; and later, while in the Paris laboratories, on the Madrid–South America high-frequency link and also on the microwave system across the English Channel. He developed the first single-sideband high-frequency radio system. He pioneered in pulse methods, inventing pulse-width and pulse-time modulation systems, and holds the patent on quantized pulse systems (pulse-code modulation). He also introduced the use of flip-flop circuits for frequency division. The multipoint gas counting tube was another of his inventions.



Alec Hurley Reeves

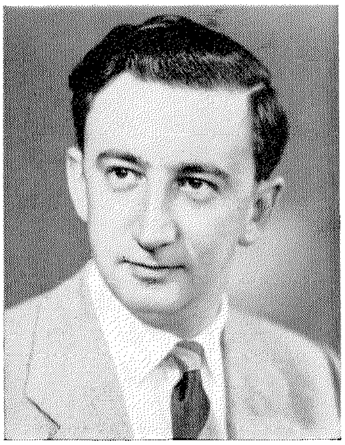
During the war, he went to the Telecommunication Research Establishment on radar work and was responsible for the development of the Oboe system, for which he received the award of the Order of the British Empire in 1945.

Mr. Reeves is now in charge of a division at Standard Telecommunication Laboratories, where he has worked on electronic switching systems. In the quest for new electronic tools, he developed the special germanium diode that is the subject of a paper in this issue.

• • •

Eugene Van Dyck was born in Antwerp in 1924. He received a degree in electrical engineering from the Université de Louvain in 1950.

He then entered the Bell Telephone Manufacturing Company in Antwerp



Eugene Van Dyck

After spending a year on a training program in the United States, he returned to Antwerp, where he has been engaged in the application of electronics to switching systems.

He is a coauthor of a paper on an electronic pulse generator for the *8A* mechanoelectronic telephone switching system.

• • •

CAMILLE WEILL was born in Lorient, France, on June 21, 1915. He received the B.S. degree and certificate of mathematics from Paris University and in 1938 the degree of electrical engineer at the Ecole Supérieure d'Electricité in Paris. During the war, he served as an officer in the French Signal Corps.

CAMILLE WEILL

In 1941, he joined the Laboratoire Central de Télécommunications and in 1951 was named head of the department in charge of electronic switching and industrial applications of electronics.

He was responsible for setting up a laboratory at the French Air Ministry, which took him out of Laboratoire Central de Télécommunications from 1946 to 1949.

Mr. Weill is a coauthor of a paper in this issue on an electronic pulse generator for the *8A* mechanoelectronic telephone switching system.

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MANUFACTURE OF GERMANIUM POWER DIODES
POLYTHENE-INSULATED VIDEO-PAIR CABLES
LOUDSPEAKING INTERCOMMUNICATING SYSTEM AT LONDON AIRPORT
NEW TYPE OF DIFFUSION CATHODE
NOVEL GAS-GAP SPEECH SWITCHING VALVE
PRECISION RADAR RANGE CALIBRATOR
CATHODE-FOLLOWER PHASE-SHIFT OSCILLATOR
TELEPHONE STATISTICS OF THE WORLD
UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM

Volume 32 **SEPTEMBER, 1955** **Number 3**

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Technical Journal of the
International Telephone and Telegraph Corporation
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Volume 32 SEPTEMBER, 1955 Number 3

CONTENTS

Manufacture of Germanium Power Diodes 147

Polythene-Insulated Video-Pair Cables 165

Loudspeaking Intercommunication System For Catering Service at London Airport 169
By J. L. Goodwin

New Type of Diffusion Cathode 172
By A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King

Novel Gas-Gap Speech Switching Valve 179
By A. H. Beck, T. M. Jackson, and J. Lytollis

Precision Radar Range Calibrator Incorporating Beacon Function 190
By R. D. Sinish

Cathode-Follower Phase-Shift Oscillator 198
By J. C. Samuels

Telephone Statistics of the World 203

United States Patents Issued to International Telephone and Telegraph System; February—April, 1955 208

Two New Equations for the Design of Filters (Correction) 178
By Milton Dishal

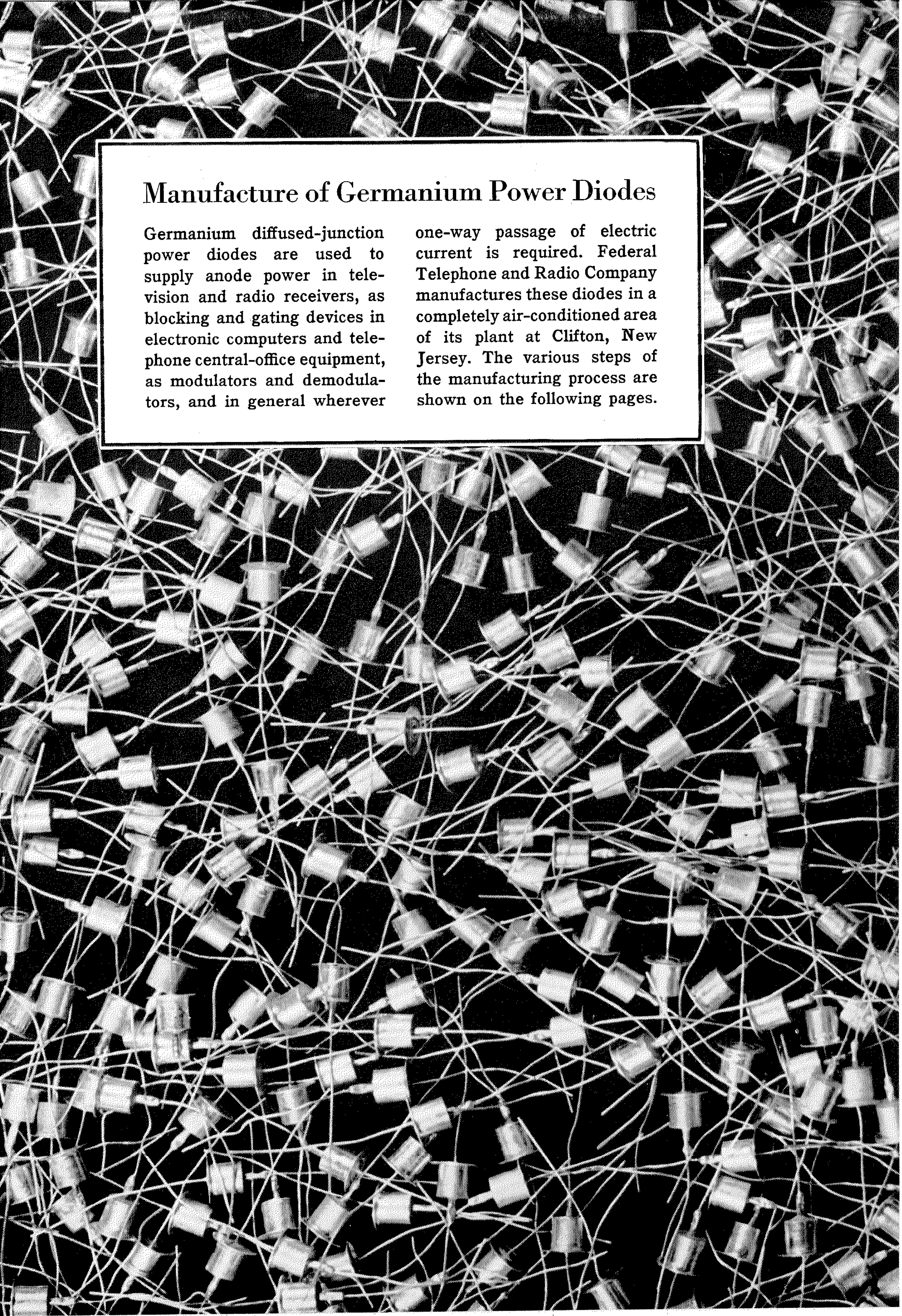
Book Reviews—
"Electronique Industrielle" 168
"Telecommunications" and "Electronics" 189

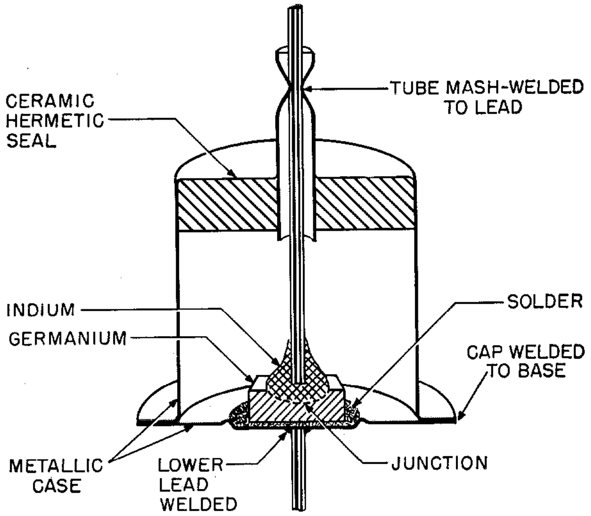
Contributors to This Issue 210

Germanium rectifier parts being assembled for heat treating.

Manufacture of Germanium Power Diodes

Germanium diffused-junction power diodes are used to supply anode power in television and radio receivers, as blocking and gating devices in electronic computers and telephone central-office equipment, as modulators and demodulators, and in general wherever one-way passage of electric current is required. Federal Telephone and Radio Company manufactures these diodes in a completely air-conditioned area of its plant at Clifton, New Jersey. The various steps of the manufacturing process are shown on the following pages.

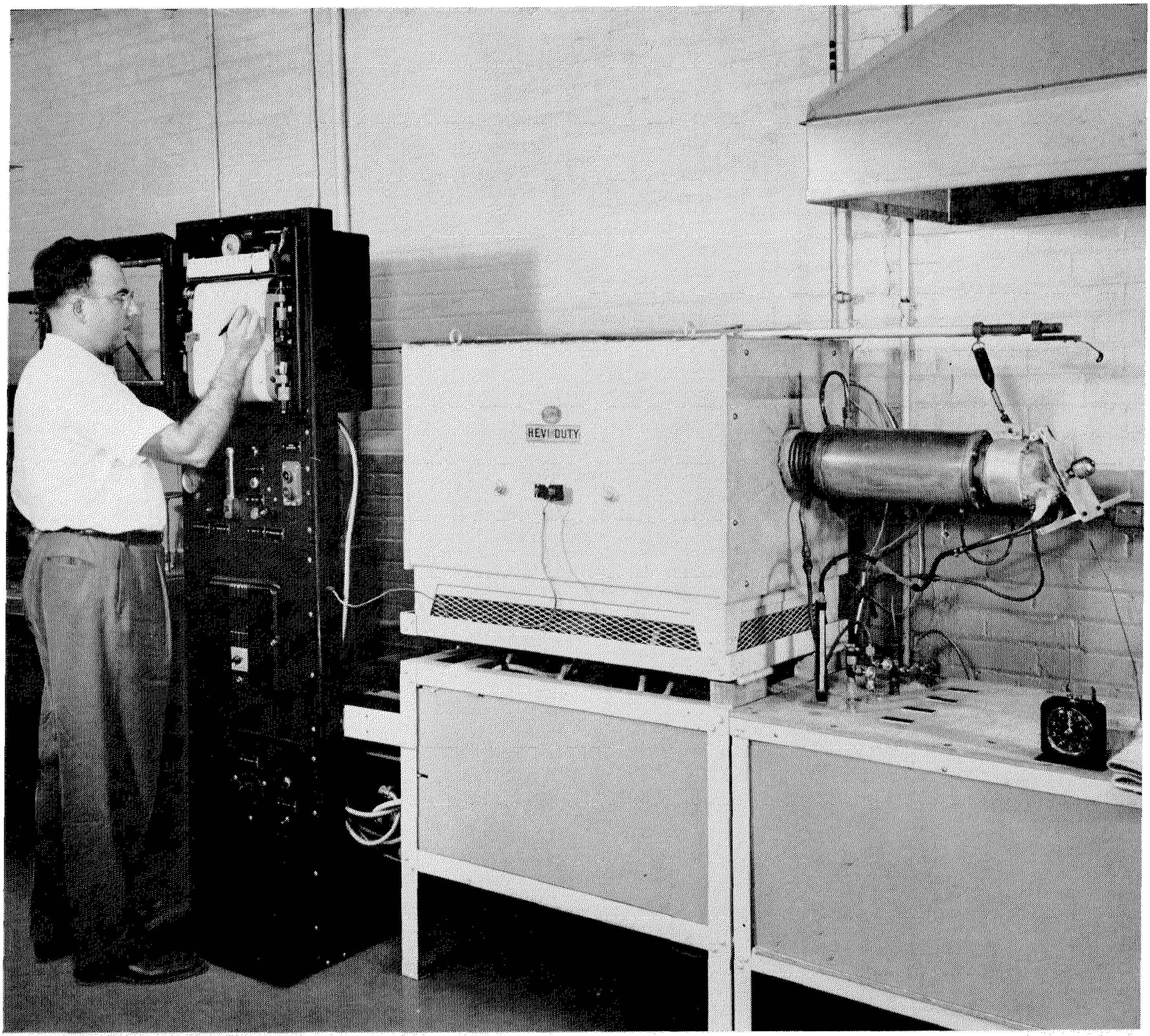




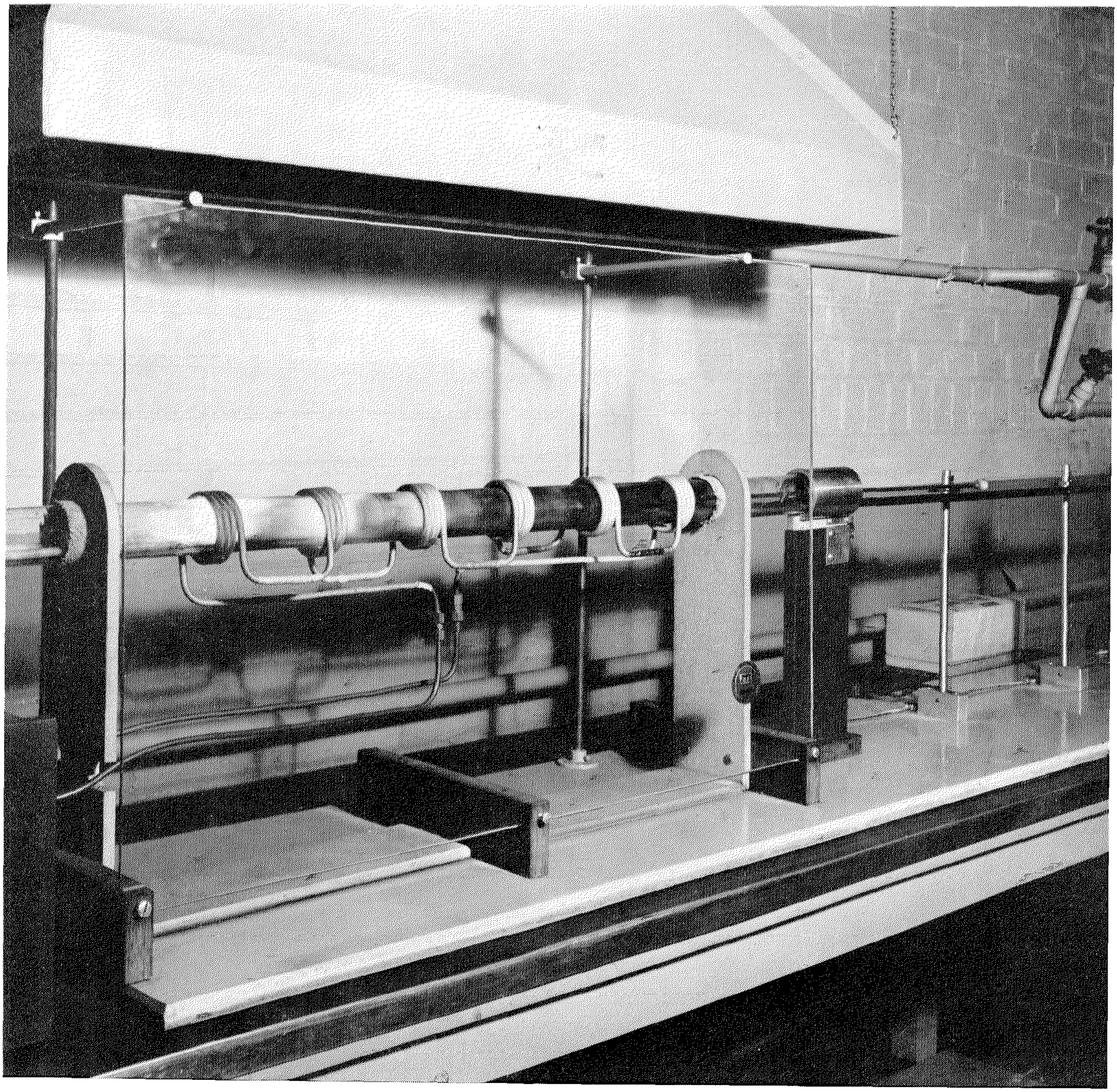
1. Cross-sectional drawing of a power diode is shown the left. The diodes pictured on the preceding page a approximately actual size.

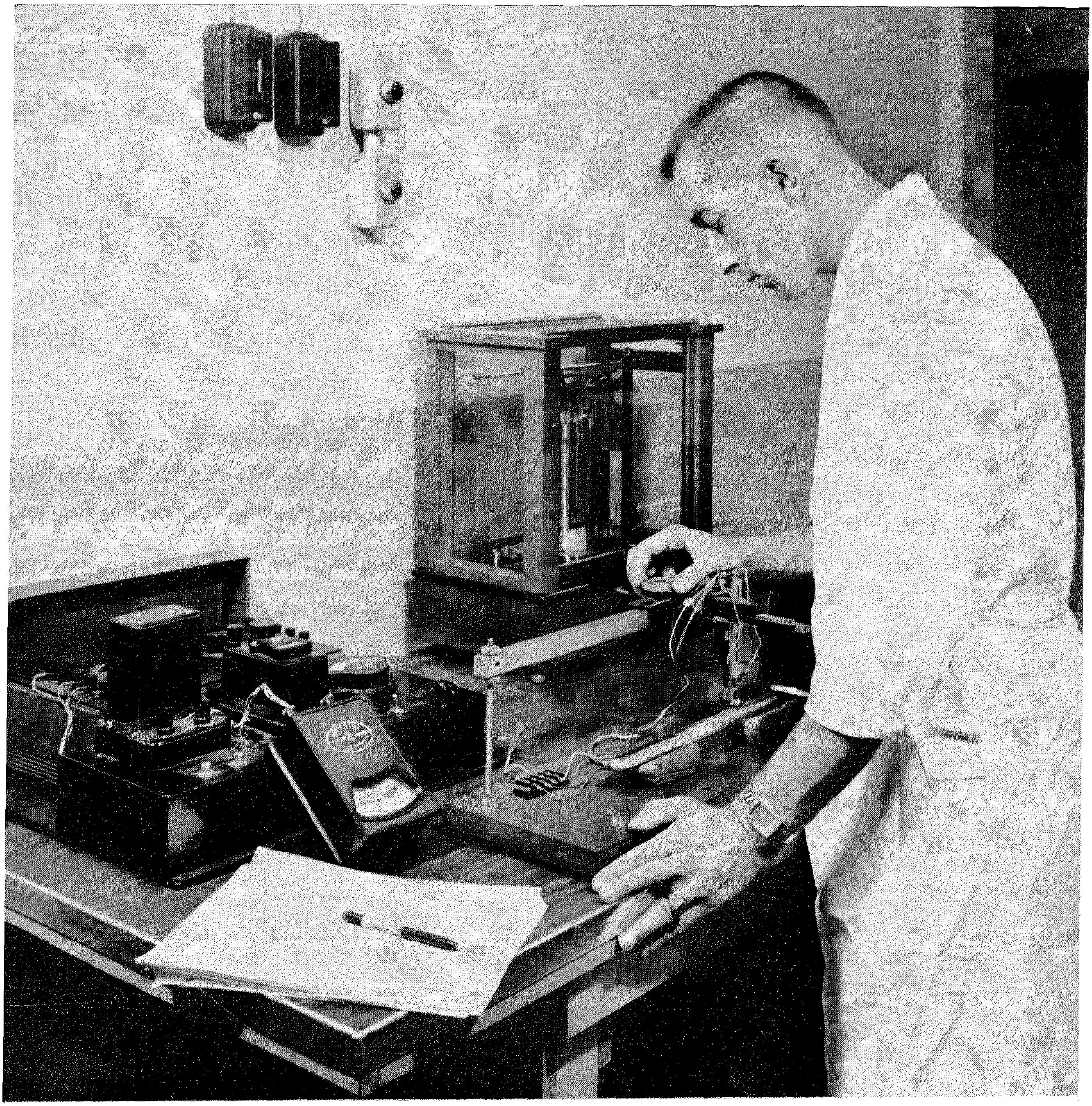
2. The first step in the production of the germaniu plates for the rectifiers is shown below. Extremely pu germanium dioxide is weighed in a graphite boat. Te perature and humidity are very critically controlled and a monitored by the instrument in the left background.

3. In this electrically operated furnace, the germanium dioxide is heated in an atmosphere of hydrogen. The dioxide is reduced to molten germanium; the oxygen combines with the hydrogen and is carried out of the furnace as water vapor. The engineer checks the recorder on the automatic temperature-controlling instrument.

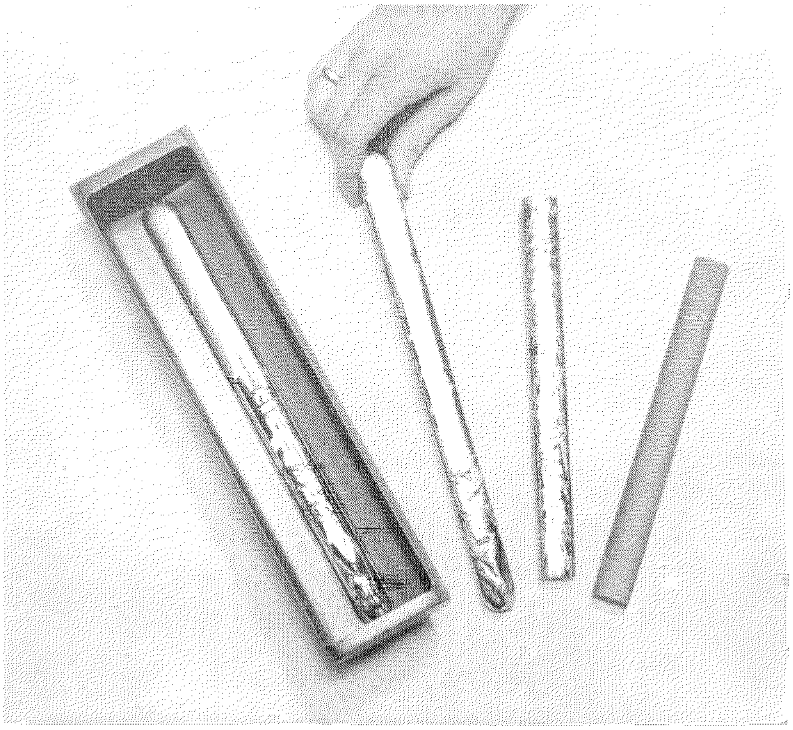


4. Successful production of germanium diodes requires very careful control of impurities in the germanium. Here is the zone-refining device. The germanium bar obtained after reduction in step 3 is placed in a graphite boat and very slowly drawn through the fused-quartz tube. The 6 coils, being energized by a high-frequency induction-heating generator, heat narrow zones of the boat to a red-hot temperature. As the boat is drawn through, each portion of the germanium bar in the boat is therefore melted and resolidified 6 times successively. Since impurities in the germanium are more soluble in molten germanium than in the solidified state, they are swept into the endmost portion of the bar, which is later cut off with a stee diamond-charged circular saw. The remaining bar consists of extremely pure germanium.



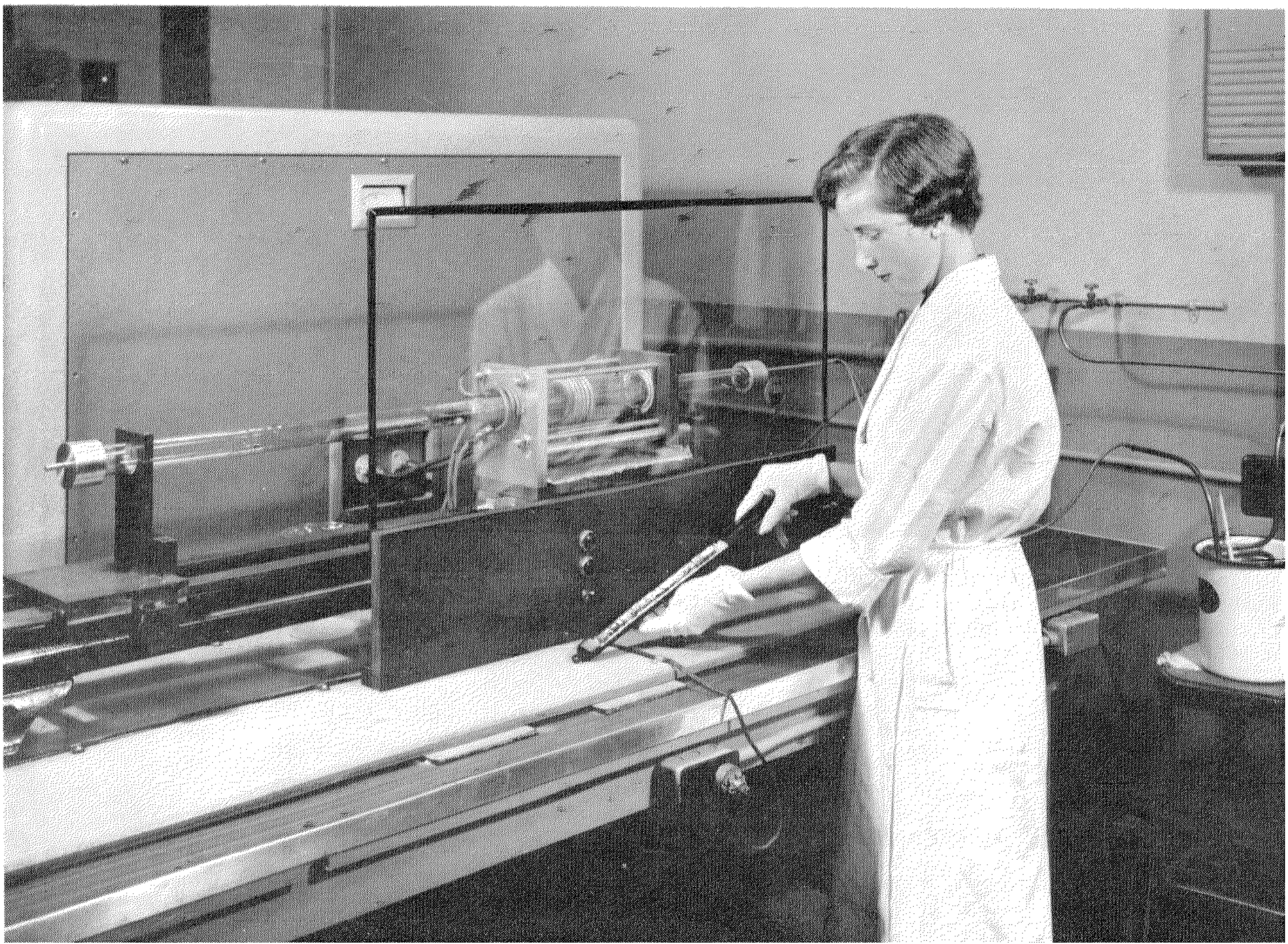


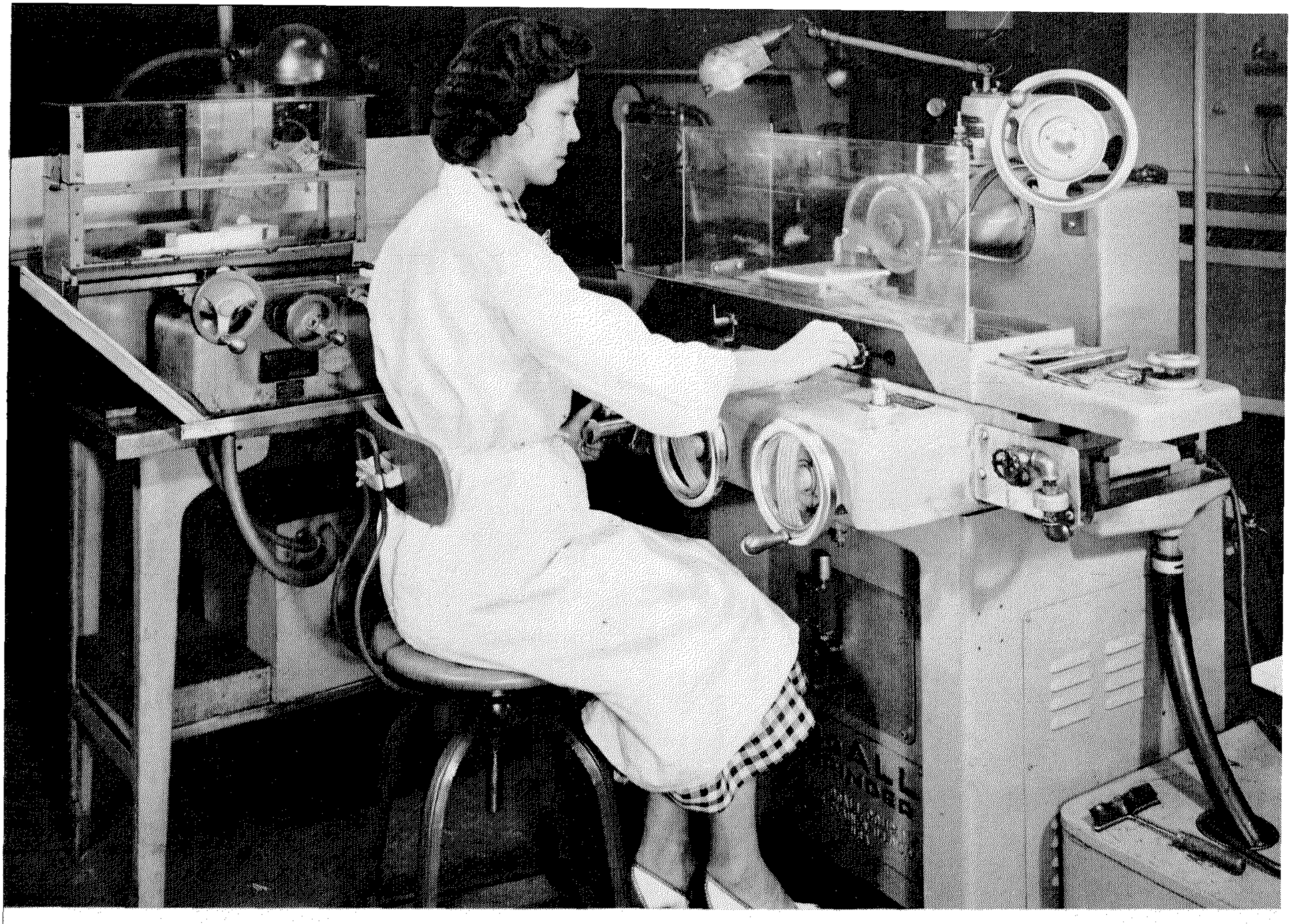
5. The electrical resistivity of germanium varies with the amount and type of impurities remaining after zone refining. Resistivity along the bar is measured accurately by means of the above device and portions having excess impurity can be removed and returned for further refining. This permits an accurate determination of how far from the end the bar must be cut off.



6. With an induction-heating device similar to that us
in zone refining, but having only one heating coil, the b
of germanium is melted in a narrow zone and as the bo
moves, it is allowed to recrystallize as a single large cryst
The worker below holds a boat used in the process. In t
front of the boat is loaded a small piece of monocrystalli
germanium. Between it and the long bar of pure ge
manium is placed a very small quantity of specially select
impurity that will give the germanium its rectifying pro
erty. Melting is started at this junction and as the melt
zone is swept along the length of the bar, the impurity
uniformly distributed throughout the bar. As the germa
ium resolidifies behind the molten zone, it conforms to t
crystalline pattern of the small piece of single-cryst
germanium; the whole bar thus becomes a single cryst

7. The stages in the production of a germanium bar a
shown at left. In the boat is a bar of germanium as reduc
from the oxide. The hand holds a reduced bar that h
been etched to bring out its polycrystalline state. Next

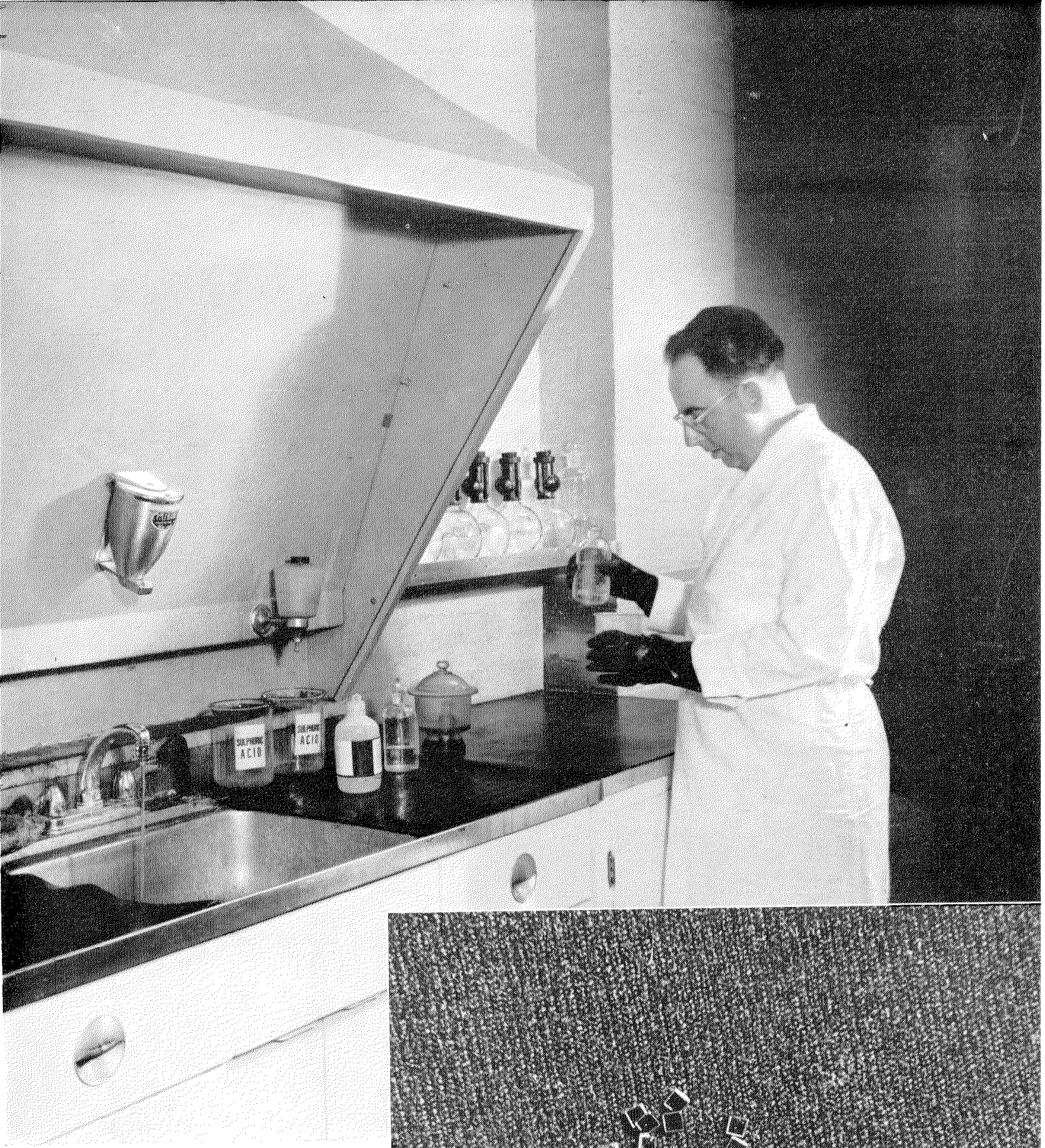


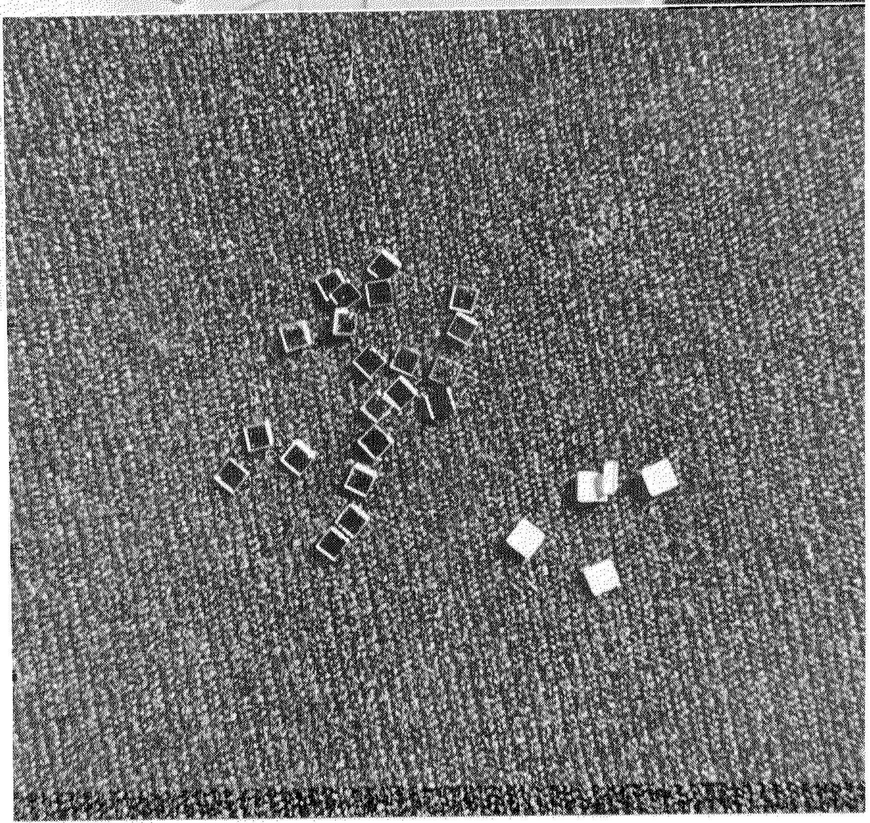


a bar after zone refining with the impure ends sawn off; this is also polycrystalline. Last is an etched bar of single-crystal germanium. After step 6, the bar is etched to remove all surface impurities and to prove its monocrystalline state; note the uniform matte surface.

8. Above are shown two machines used in sawing the germanium bar into plates that will be assembled into the diodes. At the operator's left is a saw that cuts the bar transversely into slices about 0.02-inch (0.5-millimeter) thick. These are then lapped to exact thickness and cemented with shellac onto a ceramic plate. In the machine in front of the operator, the slices are sawn into squares about 0.08-inch (2-millimeters) square. Both saws use diamond-impregnated steel disks.

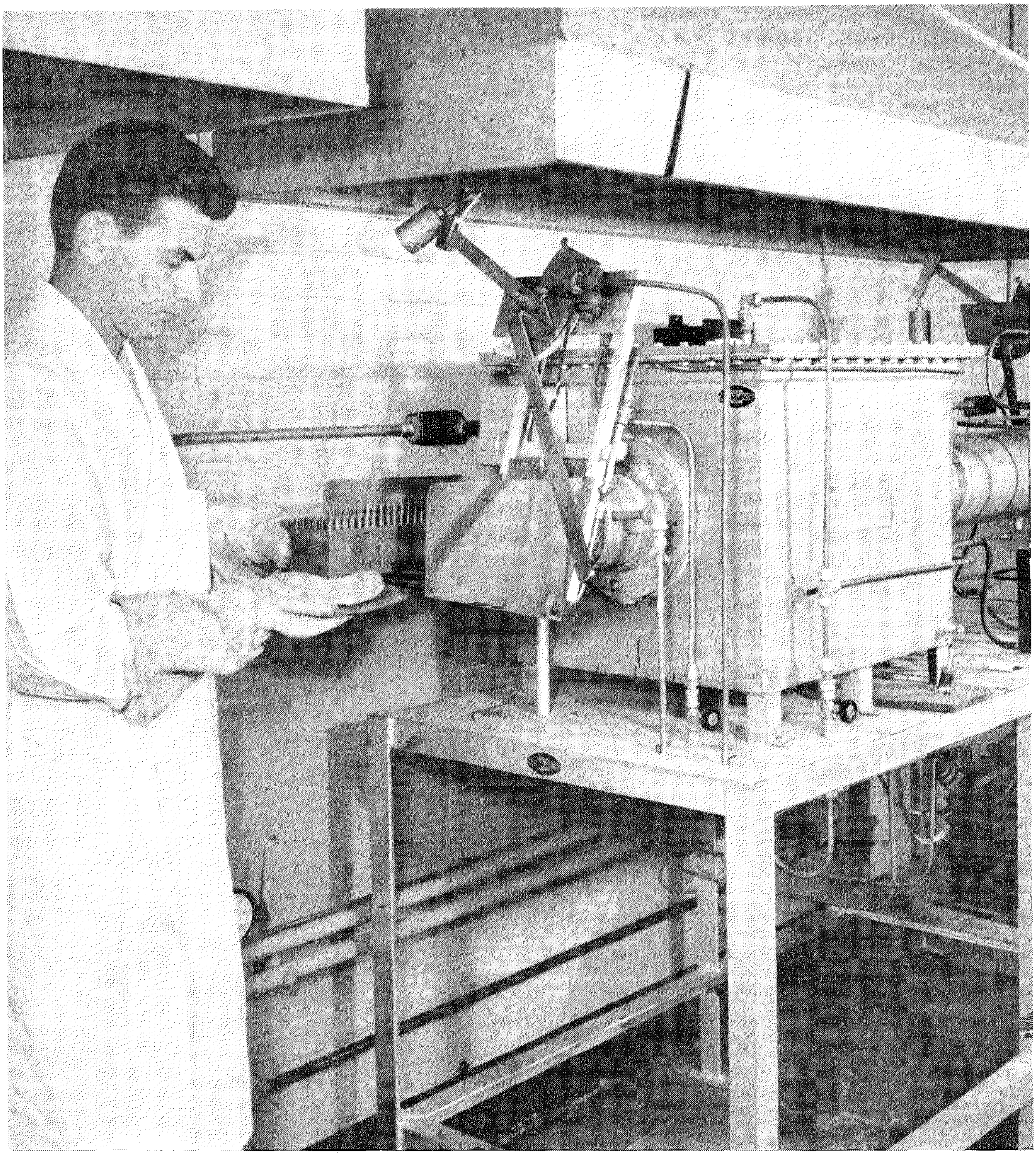
9. At the right is a piece of bar after slicing and the two plates show the slices cut into squares.



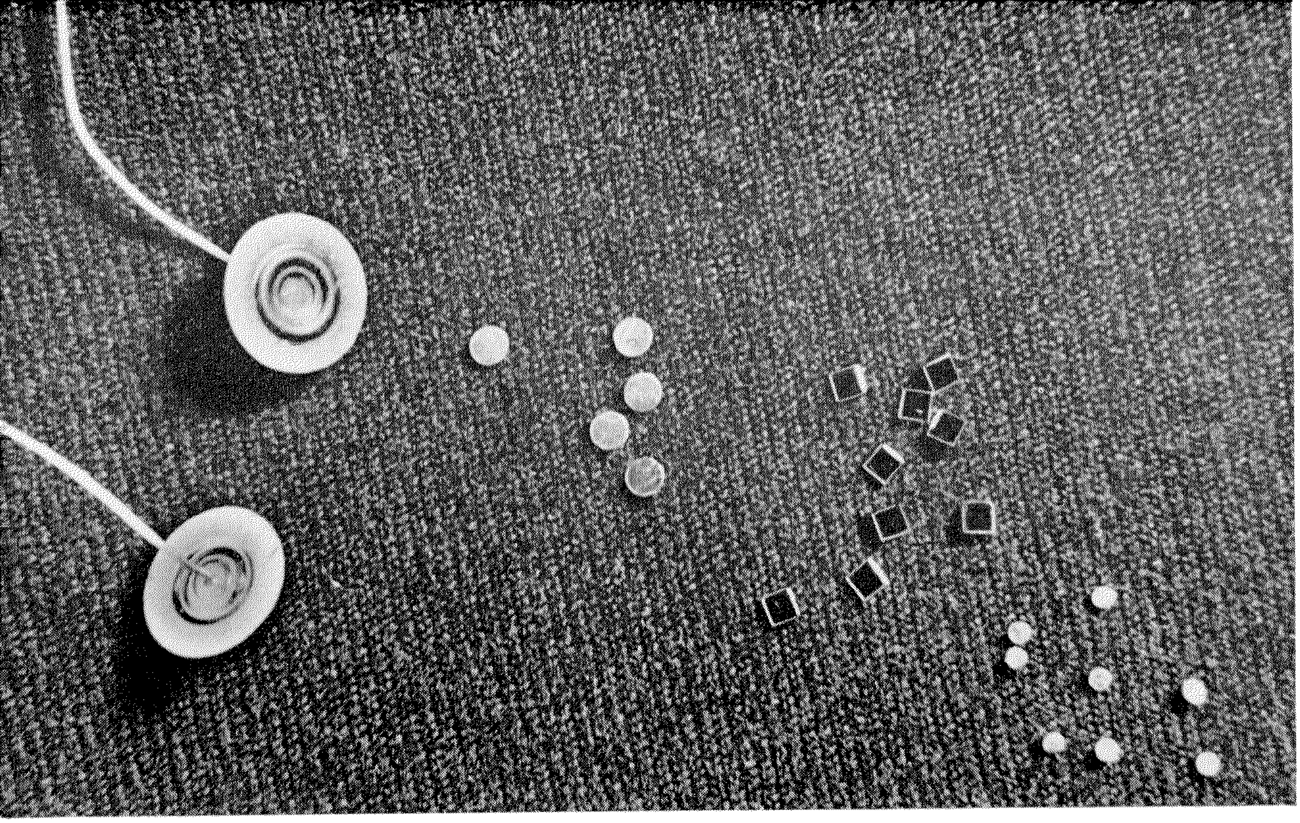


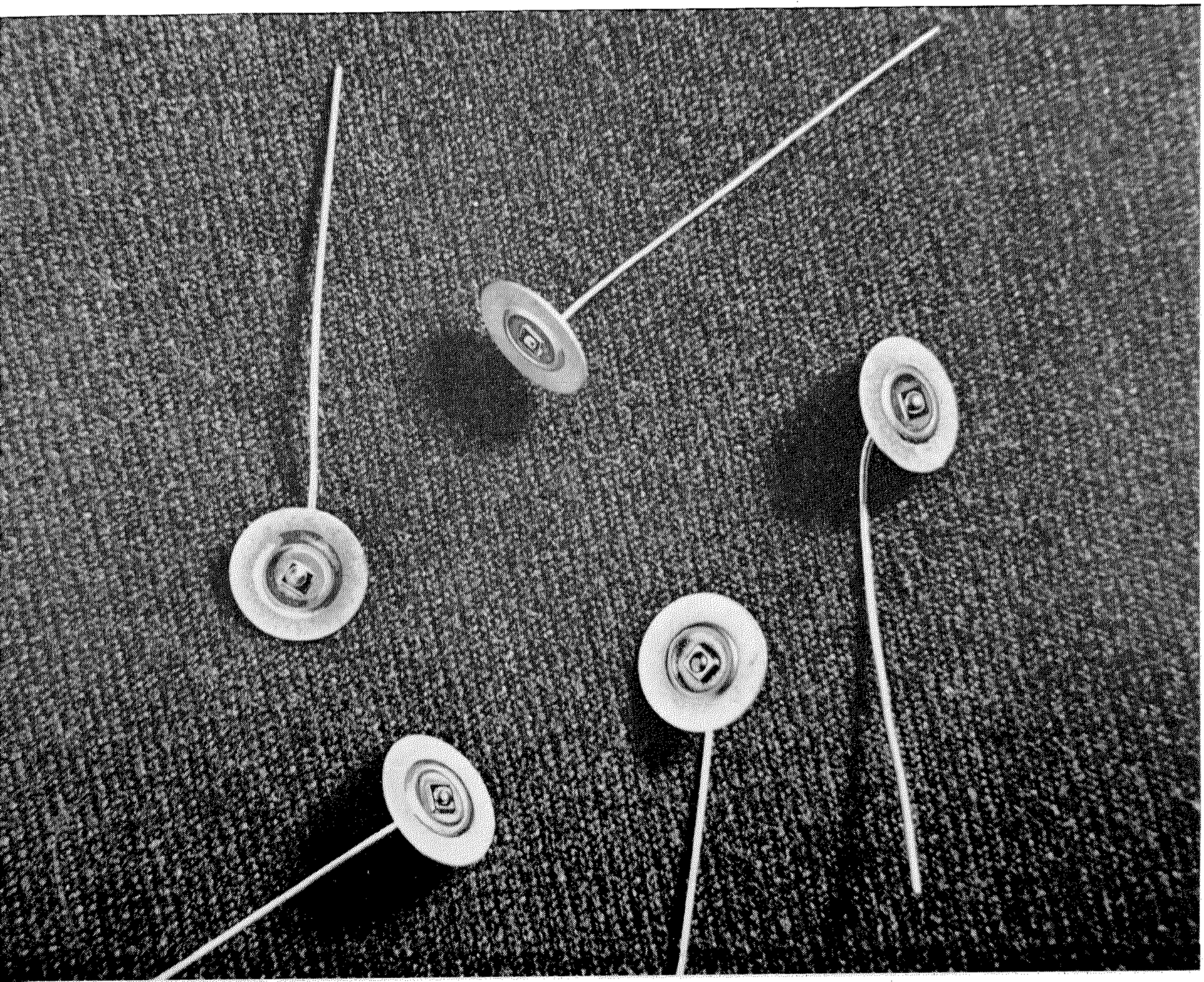
10. On the opposite page, lapping is done on a glass plate until the slices measure correct thickness on the gauge at the right.

11. After sawing, the plates are released from the ceramic plaque and are next cleaned by chemical etching as shown above. At the right, the rough bright surface of the plates are reduced to a smooth shining black by the etching.

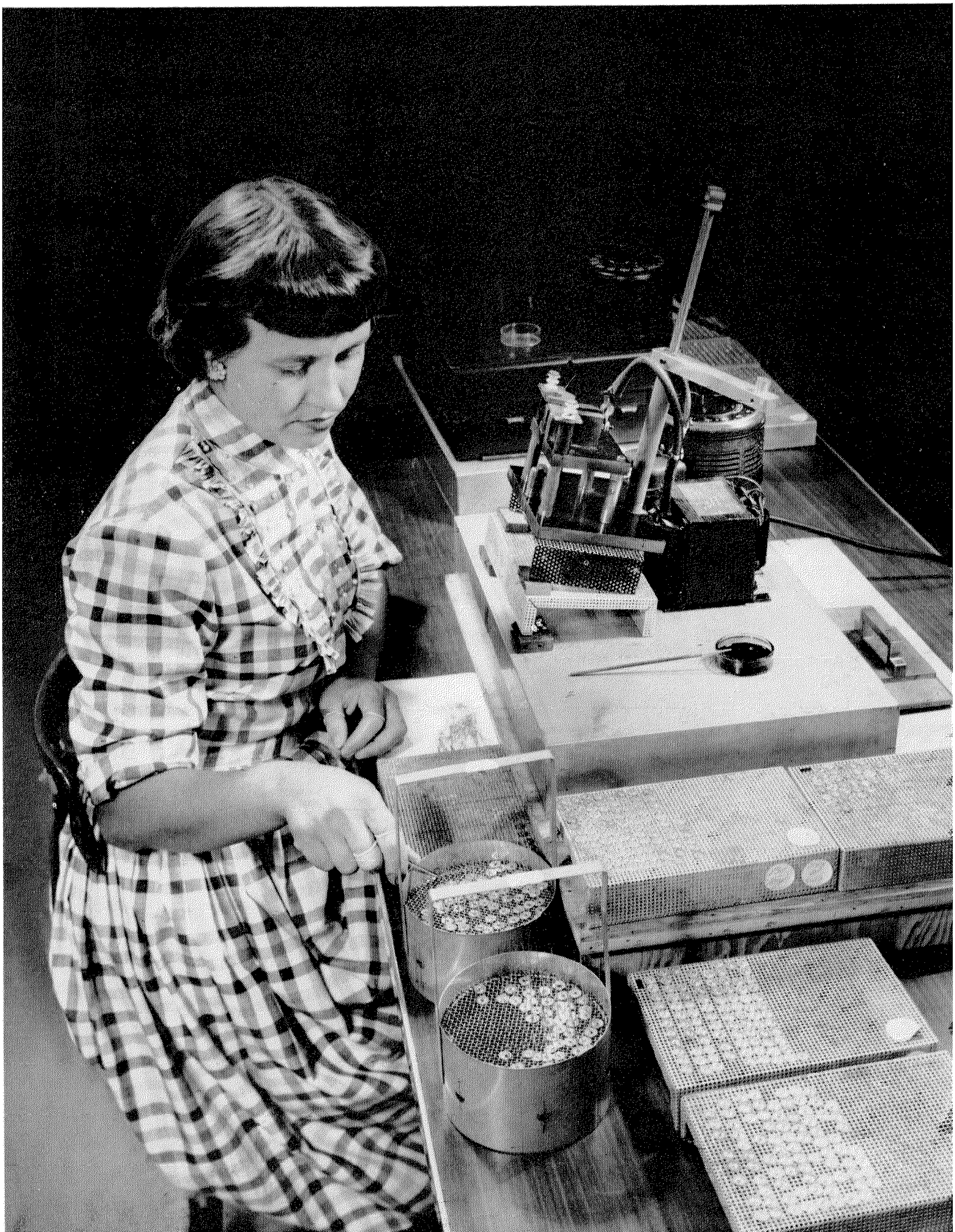


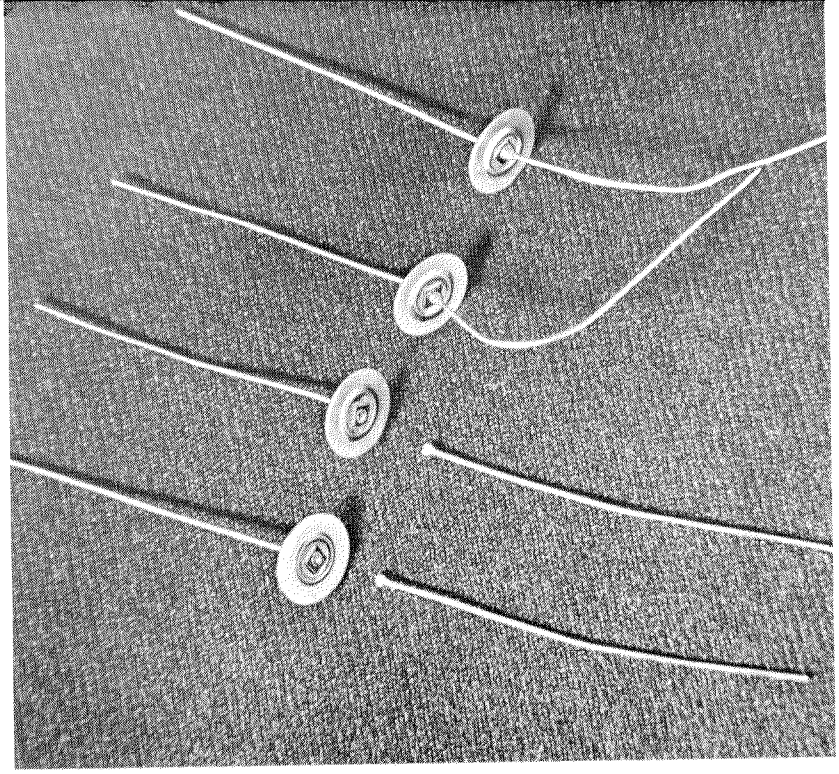
12. The frontispiece of this issue shows an operator loading a graphite boat with parts preparatory to the diffusing process. The parts are shown at the top of the facing page. From left to right are the bases, solder preforms, germanium plates, and indium pellets. Diffusion is done in a hydrogen-atmosphere electric furnace. The rectifying junction is formed at this stage, pictured above. At the bottom of the facing page are some diode bases after the diffusion stage. The small round indium pellets are partially diffused into the square germanium plates.



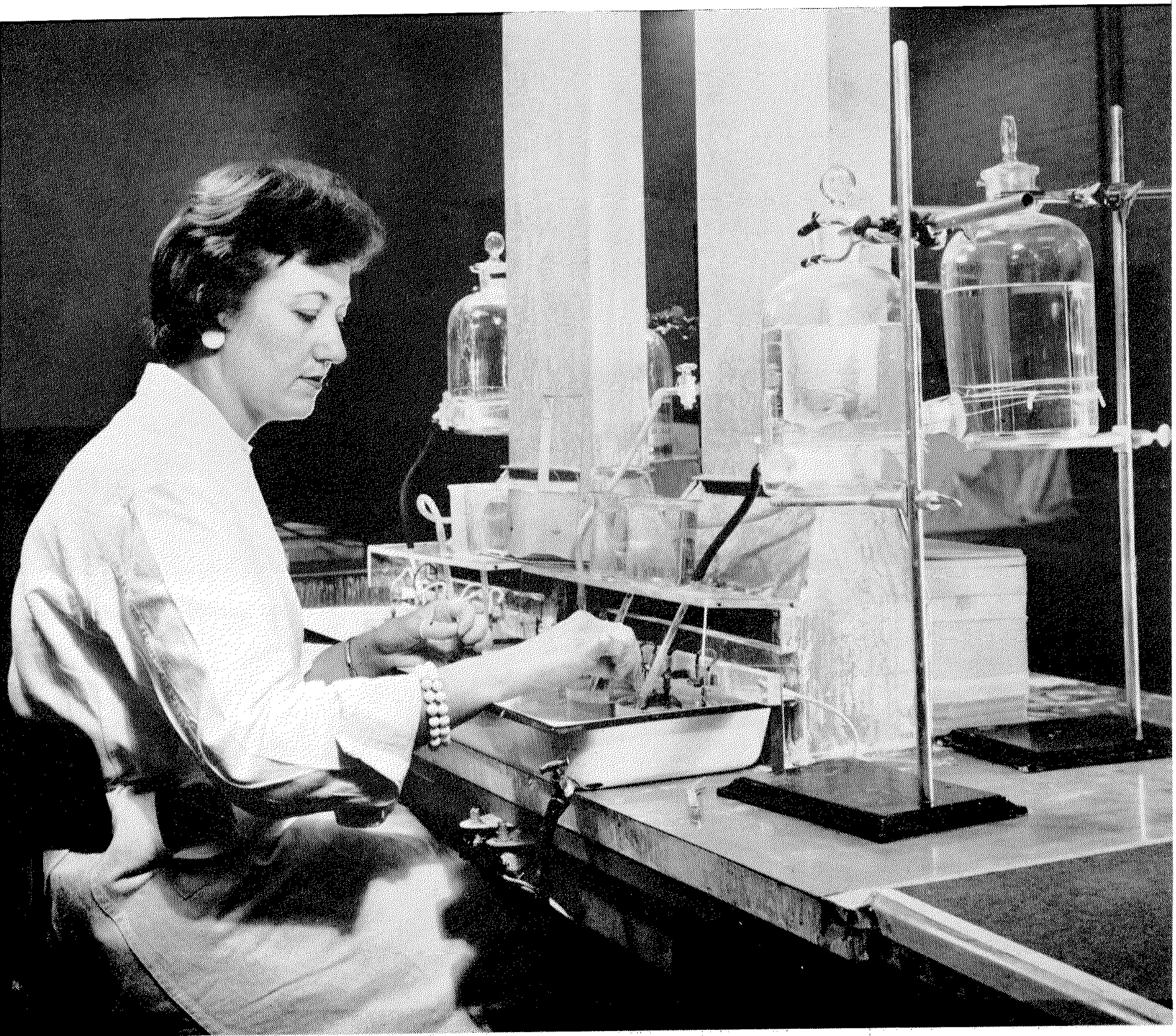


13. A pretinned nickel upper lead is fused in the indium electrode with the jig below. A preliminary measurement of forward resistance is next made to eliminate defective diodes before further operations. The process is illustrated below and at the top of the facing page.



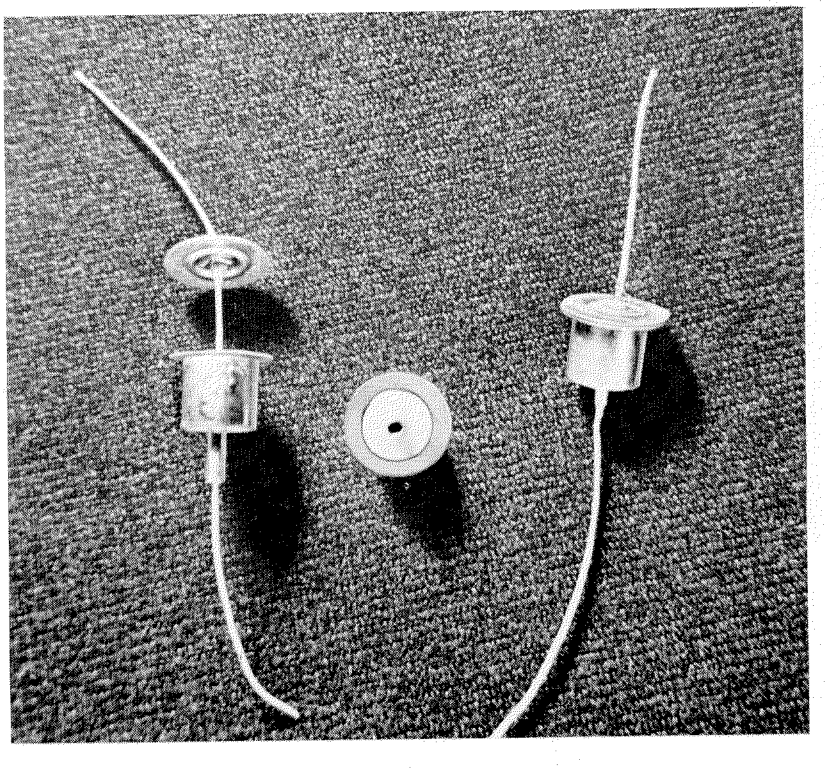


14. In the apparatus shown below, the diodes are given their final cleaning by electrolytic etching. Noxious fumes are carried off by the hood and flues.

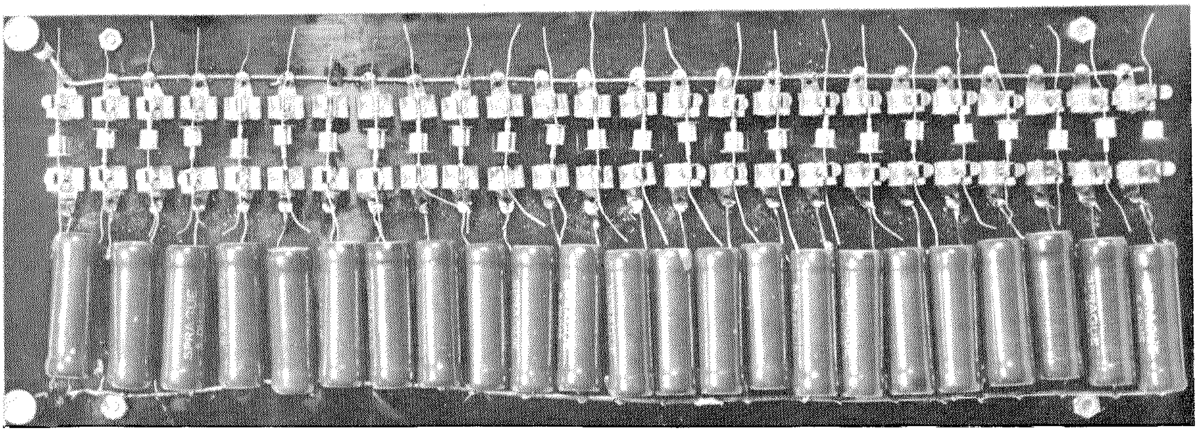


15. Characteristics of the diodes in the forward- and reverse-current directions are made at the two positions shown here.

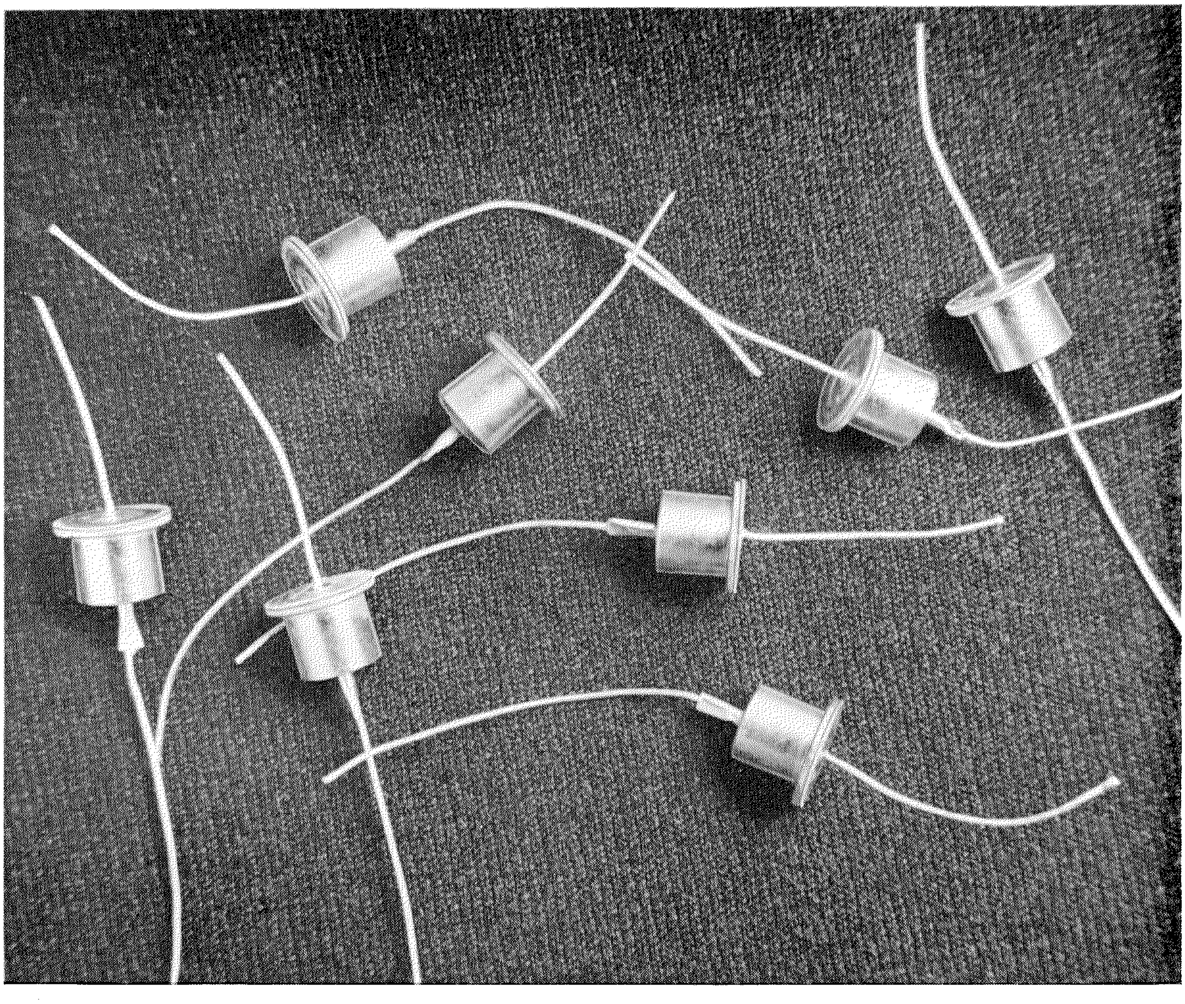
16. After testing, the active parts of the diodes are covered with a protective compound to prevent the adverse effects of moisture and other contaminants and the caps are placed over them. The caps are welded to the bases in the above machine to complete the hermetic seal. At the right is shown, left to right, the cap being slid down the upper lead wire, top view of a cap showing the insulating ceramic seal, and a welded diode. The tube is next mash-welded to the upper lead wire in a machine similar to the one above.



17. After capping, the diodes are plated for appearance and for protection against corrosion and are given several more tests. Shown here is the position for measuring the peak inverse-voltage capabilities of the diode. Two more of the testing positions are shown on the facing page. In the foreground is the preliminary forward-resistance measurement mentioned in step 13. In the rear, the diodes are measured for reverse leakage current under rated operating conditions. During this test, the diodes are placed in an oven to simulate the ambient heat to which they may be subjected during use.



18. One of the panels on which diodes are aged under operation to improve their characteristics is shown above. Finished diodes are illustrated below in about twice actual size. These will be trademarked and stamped with a type number before packaging and shipping to the customer.



Polythene-Insulated Video-Pair Cables

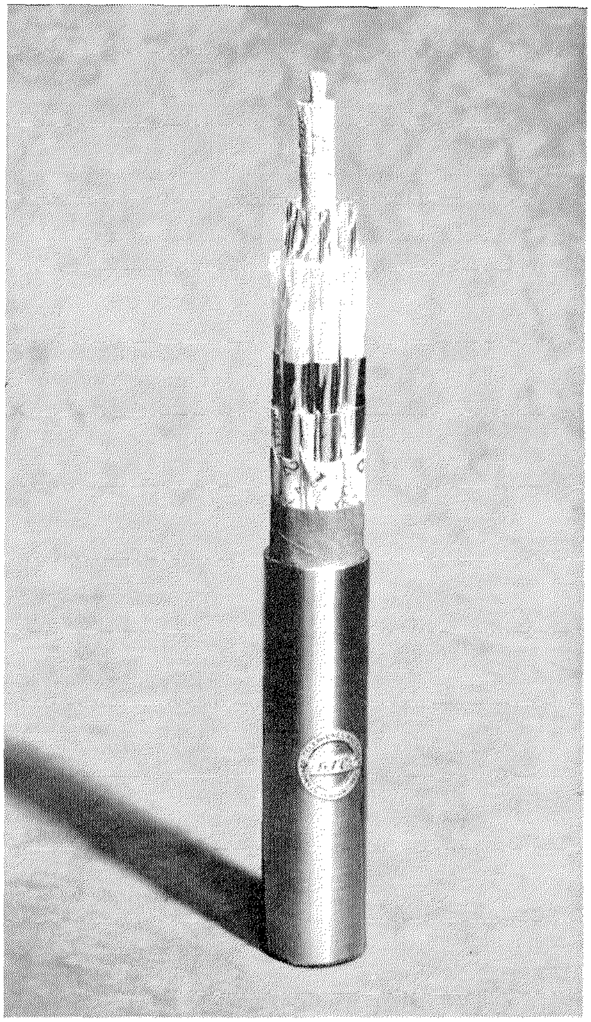
DIRECT transmission of video-frequency signals, consisting of frequency components from a few cycles per second to several megacycles per second, requires circuits that have low attenuation and good crosstalk and impedance-uniformity characteristics, but that can, if required, be connected directly to existing circuits. Conventional paper-insulated pairs are not satisfactory for video-frequency transmission over long circuits because their high shunt attenuation necessitates short spacing of the amplifier equipment and because their inadequate screening results in excessive crosstalk and interference.

Conventional coaxial cables are also unsatisfactory for direct transmission of video signals because of the ineffectiveness of the screening at low frequencies and the need for special apparatus when jointing them to paper-insulated pairs.

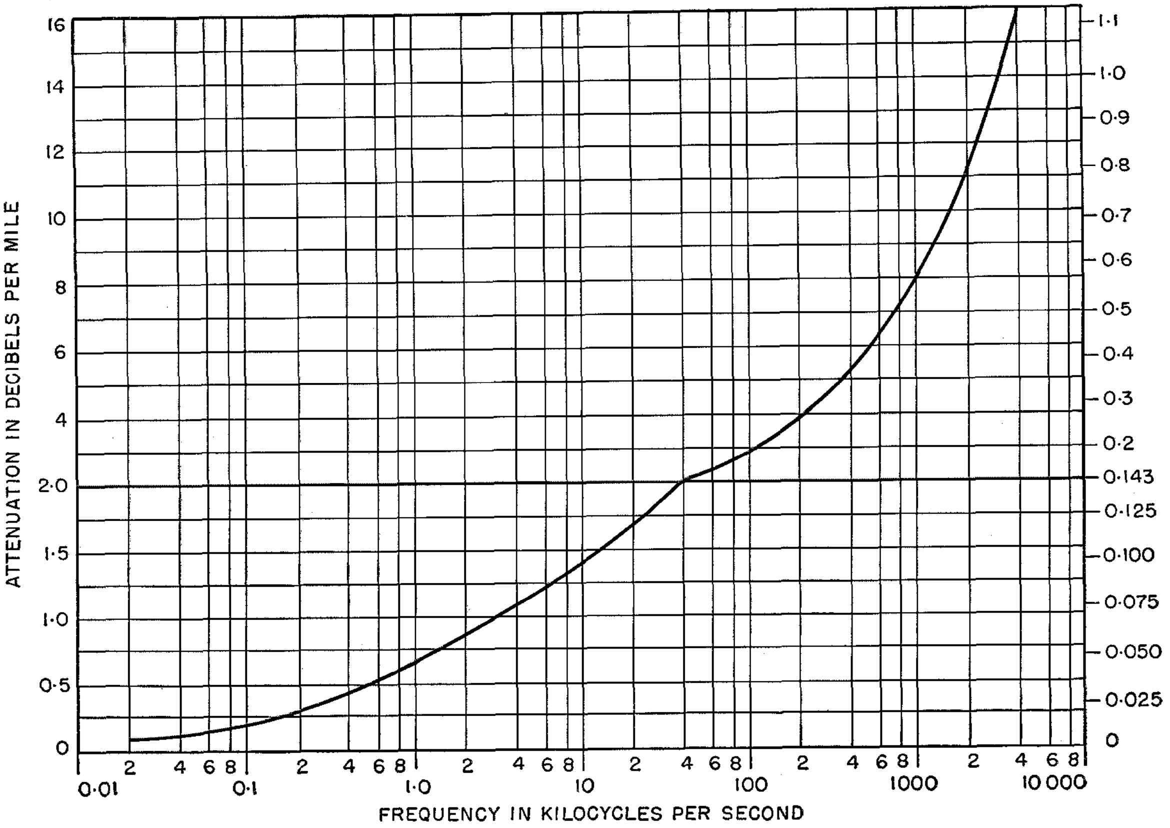
The cable required, therefore, is of a screened-pair construction with a dielectric of low permittivity and power factor, operating from about 20 cycles per second to 5 megacycles per second. A screened video pair with polythene string and tape replacing the paper string and tape of the conventional paper-insulated-conductor type is now manufactured by Standard Telephones and Cables, Limited, London, to satisfy all of these requirements. The design of this pair has the advantage that it can be manufactured on existing paper-insulated-cable plant with only minor modification.

The insulation of the new pair consists of a polythene string applied as an open helix on the conductor, followed by a closed helical wrapping of polythene tape, both applications being made in one operation on a paper-insulating machine. Two such insulated conductors are then twisted together with two fillers consisting of twisted polythene strings in the outer interstices to form a round unit. The pair so formed is lapped with polythene tape, then screened, first with a longitudinally applied copper tape, then with a helically applied copper tape; and finally lapped with two paper tapes, the outer one of which is marked with an identification number. The overall diameter of the pair is approximately 9 millimetres (0·35 inch) for a 1·27-millimetre (0·05-inch) conductor, giving optimum spacing between conductors and screen for minimum attenuation.

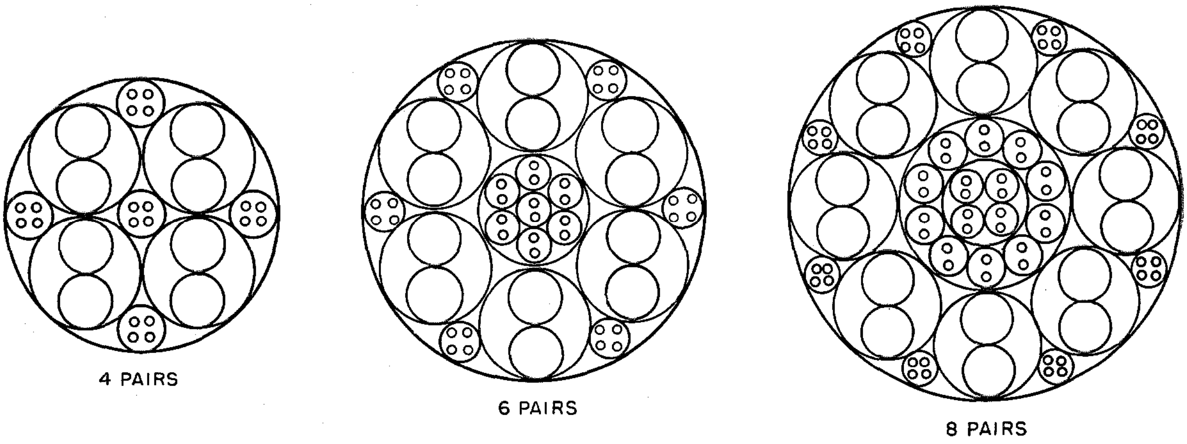
The above design keeps crosstalk to the low values required for the transmission of video-frequency signals. The longitudinal crosstalk due to electric-field coupling between the wires of the pair and their screen depends on the difference in capacitance between each wire and the screen (side-to-earth unbalance) and on the coupling to the pair of a voltage induced on the inner



Cable containing 8 polythene-insulated video-frequency pairs. There are also 8 star quads and 14 pairs in the interstices.



Attenuation of video-pair cable versus frequency; temperature was 60 degrees fahrenheit (15 degrees centigrade). Note expansion of scale below 2 decibels per mile.



4 PAIRS

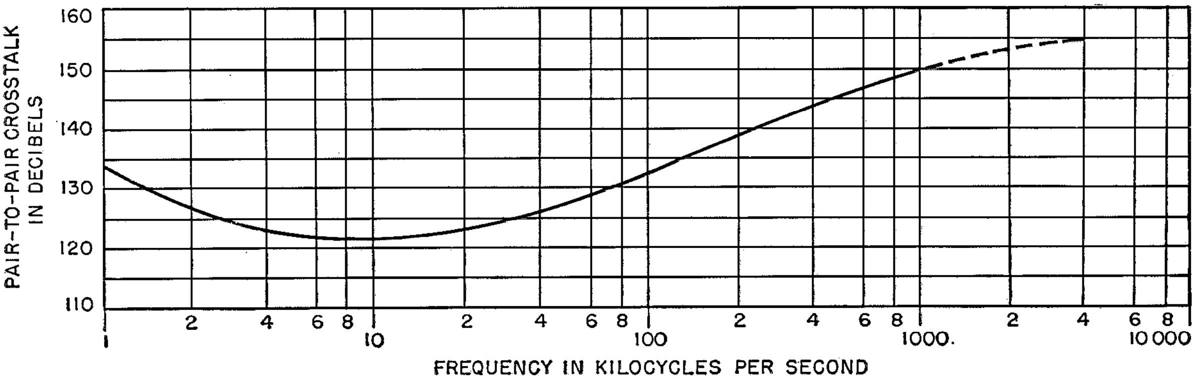
6 PAIRS

8 PAIRS

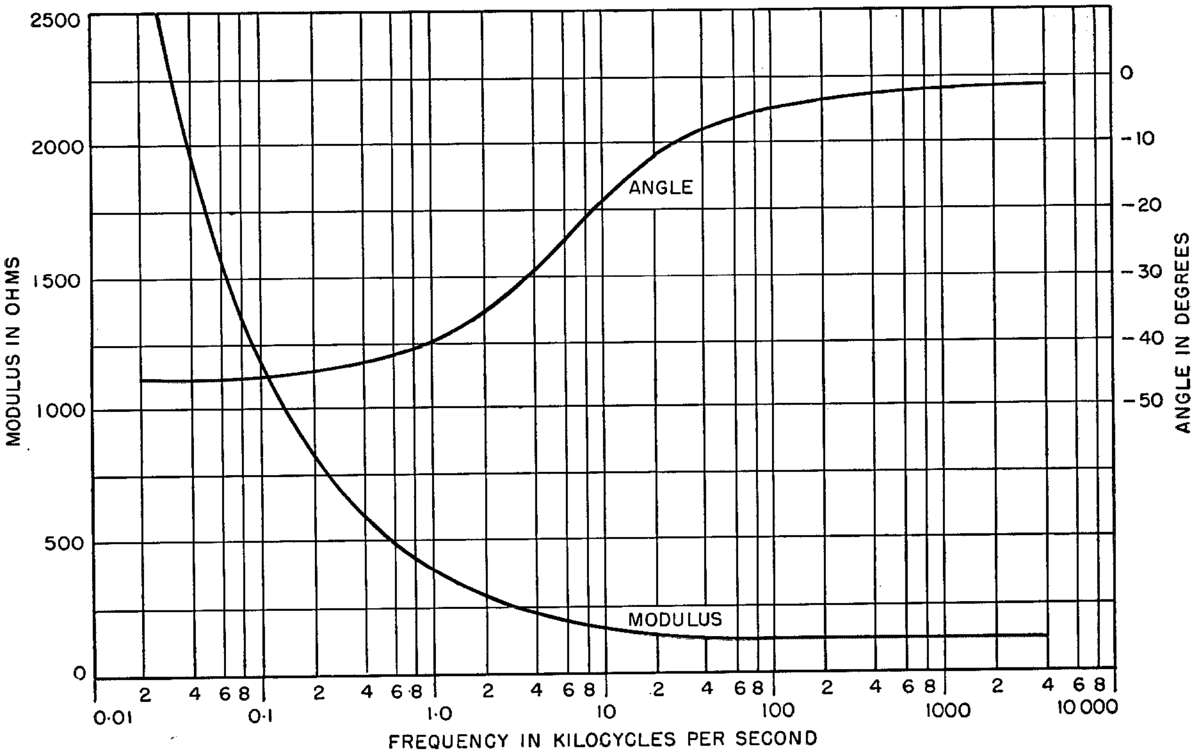
Three typical cable designs incorporating polythene-insulated video pairs. The 4-pair unit has a star quad in the centre and 4 outer-interstice star quads, all with 0·63-millimetre (0 025-inch) conductors; overall diameter is 22 millimetres (0·87 inch) and the weight is 2150 kilograms per kilometre (7630 pounds per mile). The 6-pair unit has 7 pairs in the centre with 0·9-millimetre (0·035-inch) conductors and 6 star quads with 0·63-millimetre (0·025-inch) conductors in the outer interstices. Diameter is 27 millimetres (1·06 inches) and weight, 2940 kilograms per kilometre (10 434 pounds per mile). The 8-pair cable has 14 pairs in the centre with 0·9-millimetre (0 035-inch) conductors and 8 star quads with 0·63-millimetre (0·025-inch) conductors in the outer interstices; diameter is 32·5 millimetres (1·28 inches) and weight is 3850 kilograms per kilometre (13 664 pounds per mile).

surface of the screen from an interference current on the outer surface. This voltage depends on the surface impedance of the screen and for this reason the first copper tape of the video pair is applied longitudinally to give less surface impedance than is obtained from a spiral lapping. With the latter, at the higher frequencies, the current is affected by the contact resistance between turns. The side-to-earth unbalance is kept at a minimum by careful control of the materials comprising each pair and by uniformity of manufacture. Transverse crosstalk due to magnetic coupling between the pairs is reduced by suitable selection of the pair twists and by the adequacy of the shielding.

An alternative insulation for the video pair

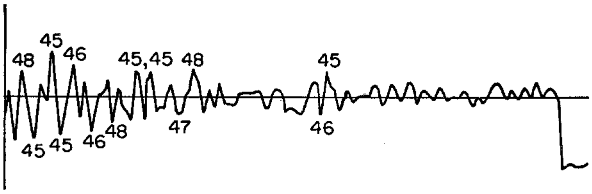


Typical pair-to-pair crosstalk curves plotted against frequency.



Impedance-versus-frequency curve for video-frequency cable.

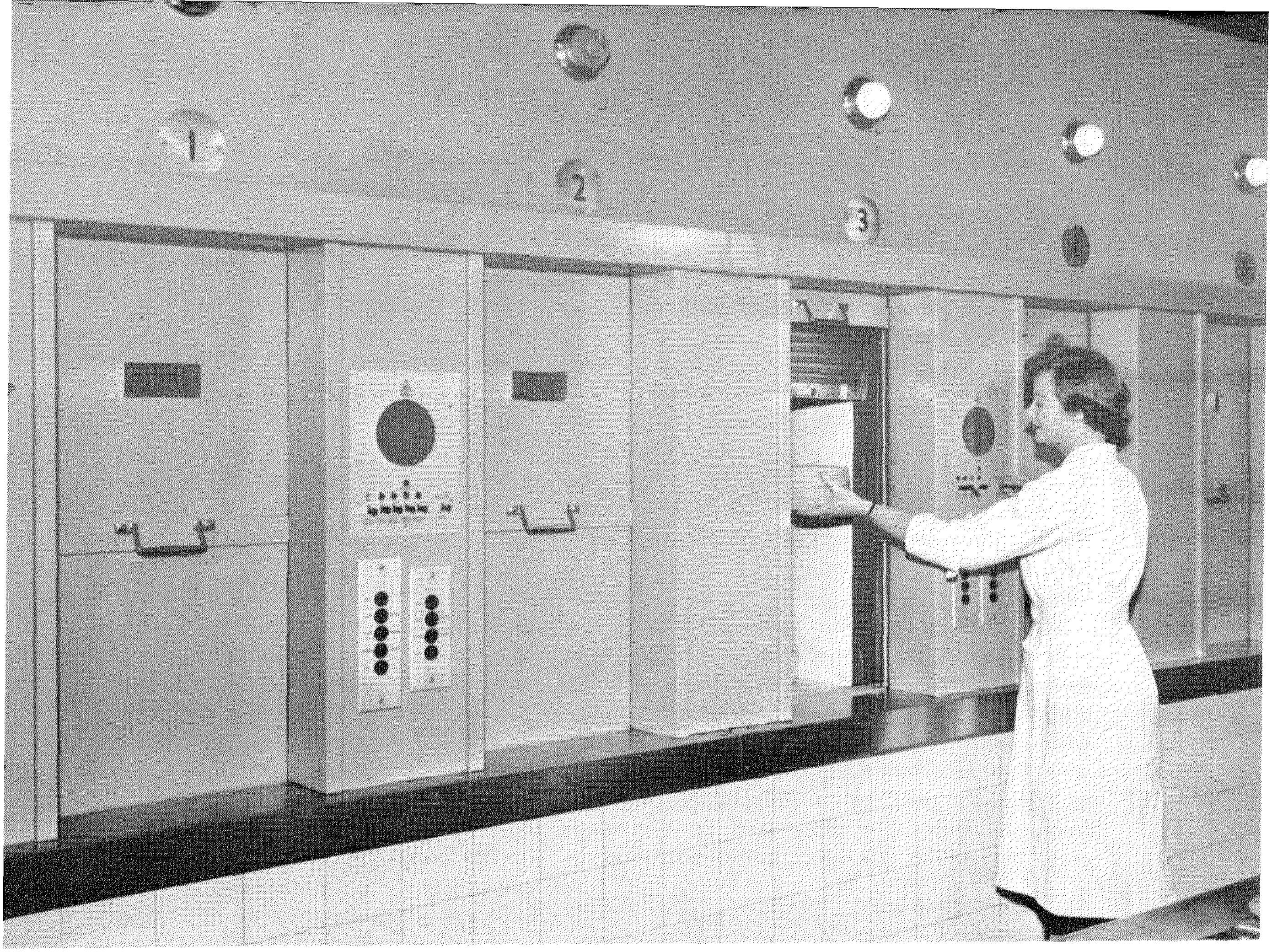
is expanded polythene. A dielectric of approximately 50-per-cent polythene and 50-per-cent air gives approximately the same attenuation as the polythene string-and-tape insulation.



Pulse-reflection oscillogram illustrating impedance uniformity of the vido-pair cable. The transmitted pulse length was 0·05 microsecond. Cable length was 500 metres (547 yards). The values of the major line reflections (corrected for line attenuation) are indicated in decibels below transmitted-power level.

Any reasonable number of video pairs may be stranded together with their associated control and supervisory circuits to form a composite cable around which paper-insulated pairs or quads may be stranded if required. The cable is then lead-sheathed and armoured in the normal manner.

The accompanying illustrations give the major characteristics of this new type of cable construction. The cable indicated in the curves had 1·27-millimetre (0·05-inch) conductors.



Completed installation showing lifts with intercommunication panels immediately above the lift-operation controls.

Loudspeaker Intercommunication System For Catering Service at London Airport

By J. L. GOODWIN

Standard Telephones and Cables, Limited; London, England

THE South East Face building of the new Central Terminal was designed to handle the bulk of the passenger traffic at that London airport and is completely equipped with full catering facilities. The planning of a modern kitchen invariably divides it into a number of areas devoted individually to such services as food preparation, wine store, and wash-up for tableware. A flexible communication system is required not only among these areas but must include the service lifts or elevators that are part of the food-handling system.

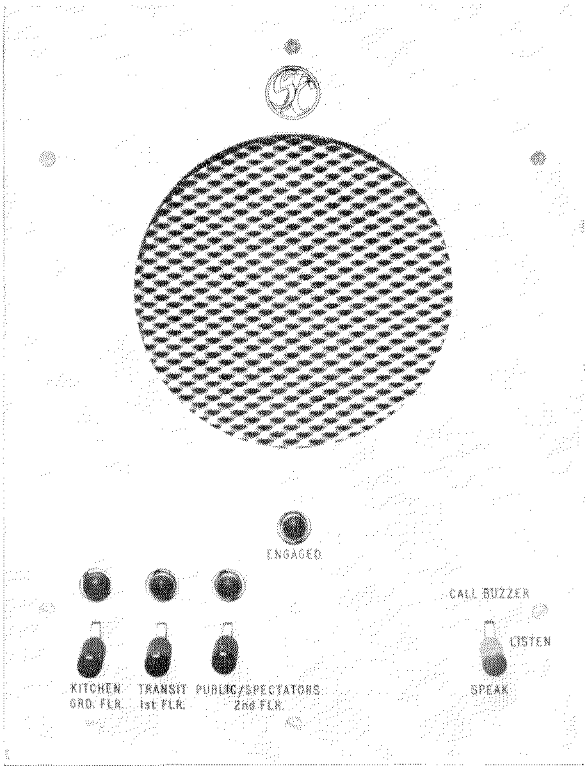
In this particular installation, a loudspeaking telephone having a centralized master station and a number of extensions would not do, for it would not be practicable to nominate one area as the master to which all calls would have to be made. A requirement was that any pair of side stations should be able to talk to one another.

The installation includes a bank of 10 food service lifts and 2 others, all of which are unattended by operators and do not carry personnel. The lifts serve a maximum of 4 floors and in some instances there is access to the car from two opposite sides. An example of this is on the first floor where the transit restaurant is on one side and the wash-up room is on the other.

To cover this installation, 7 loudspeaking telephone systems were required, each system having from 2 to 6 extensions. The design was based on the familiar "talk–listen" principle with the basic circuit and terminals the same for each system. This standardization resulted in manufacturing and installation economies and in simplified maintenance.

In each system, a common amplifier permits transmission between 2 of the extensions at a time. A small loudspeaker at each station also serves as the microphone and relays are actuated to connect the amplifier in the proper direction for speech transmission from the station when the telephone-type key is in the "speak" position. When the key is released, it returns to the "listen" position.

The input and output terminals of the amplifier appear on opposite sides of the same spring set, and whilst this would generally be undesirable when using a high-gain amplifier with a wide frequency response, it is practicable with the amplifiers used for these systems. The gain is of the order of 70 decibels and the high-frequency response is attenuated. In more elaborate systems it has been found necessary to utilize special switches that usually result in the signal passing through a greater number of contacts; if an attenuated high-frequency response can be accepted, that is the better solution to adopt.



Typical control panel for a 4-extension system.

For the convenience of working over long lines and of finding a suitable impedance for transmission and reception, the amplifier is designed to work between impedances of 600 ohms. The transducer units of a nominal 3-ohm impedance are fitted with transformers to convert their impedance to 600 ohms.

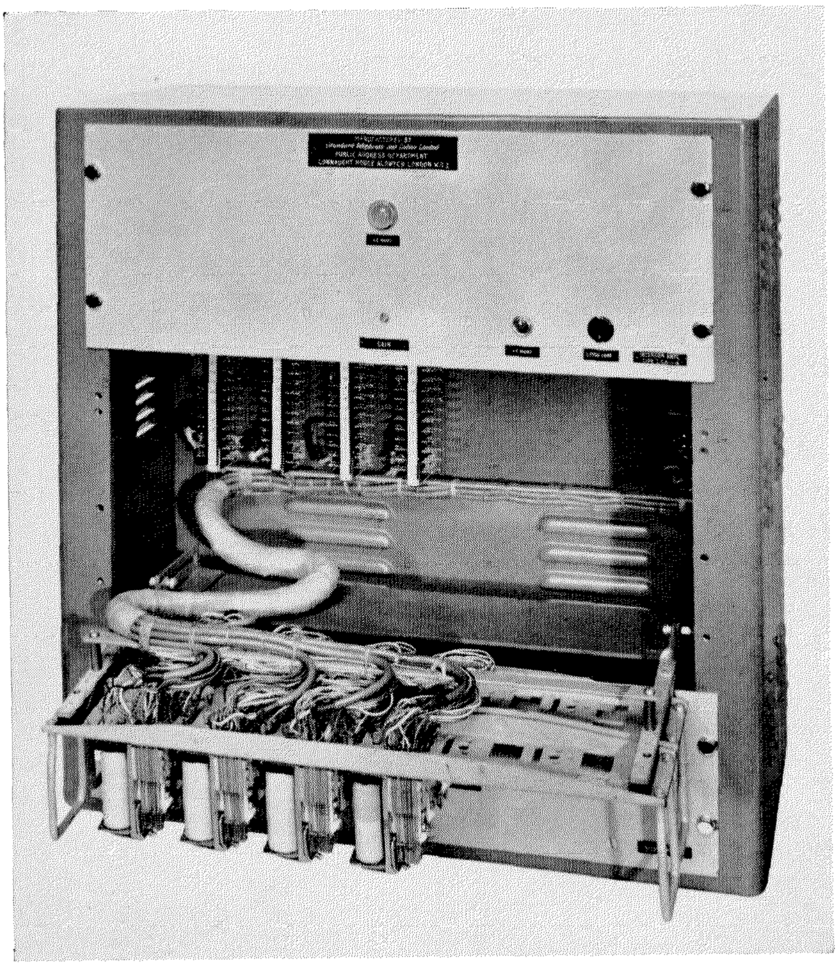
Calling may be effected by using voice or by operating the "Talk/Listen" key in the upward direction to the position "Call Buzzer." Contacts on a relay operate the buzzer and connect its pulsating output through a suitable attenuator to the input of the amplifier. By this means it is possible to employ a central buzzer, using the amplifier to produce the required volume, instead of local buzzers at each extension sounding from within the panel housing.

In common with other intercommunication systems, the extension under call is controlled by the party in the calling position and the called party must therefore pause momentarily before replying to allow time for the throwing of the switch by the caller.

Very simple lamp circuits are used to indicate to the initiator of a call that the system has been seized and further circuits indicate to other parties that the system is engaged.

Bass attenuation has been applied to overcome any low-frequency resonances associated with the loudspeaker unit and its enclosure, and to reduce the effect of the acoustic feedback that can occur through the lift shaft from one extension to another. High-frequency attenuation is included to minimize the effects of response peaks in producing excessive sibilant tones. In addition to this, a decreasing gain with ascending frequency is desirable to ensure that the switched system is stable under conditions of maximum sensitivity.

The power output had to be sufficient to accept without distortion an input from a person speaking loudly or even shouting. A further consideration was that this system be capable of connecting to more than one extension at a time. Under normal operating conditions and for maximum performance, only two extensions are interconnected but it must be possible for the initiator of a call to select more than one extension. In the speaking condition, the amplifier output will be divided among the total number of extensions engaged by the initiator. In the listening condition, all of the called transducers become parallel-connected microphones with a resulting loss in sensitivity and an increased background noise. To meet these requirements a maximum output of 1·5 watts is available and trials have shown this to be a suitable value.



Apparatus cubicle with centre panel removed allowing the relay frames to be hinged forward for servicing and exposure of the terminal strip.

The amplifier, relays, and direct-current supply unit are assembled in a sheet-steel cubicle approximately 22 inches (56 centimetres) square and 10·5 inches (27 centimetres) deep.

The extension units are panel mounted and fitted into flush wall boxes. The panels may be mounted in any suitable enclosure. The flush wall boxes are fixed behind the lift architraves. This is convenient as the controls for the lifts are mounted in a similar manner just beneath the intercommunication panels, presenting a neat and pleasing appearance to these control units.

New Type of Diffusion Cathode*

By A. H. BECK, A. D. BRISBANE, A. B. CUTTING, and G. KING

Standard Telecommunication Laboratories, Limited; London, England

1. *Introduction*

VALVES today are required for telecommunication channels at higher and higher frequencies so the importance of cathodes yielding greater current density with reasonable life under continuous operation increases. Thus in recent years various types of cathodes have been produced which sacrifice emission efficiency for better continuous-service operational characteristics. Notable examples of these are the Lemmens cathode[1,2], the modified *L*-cathode using barium aluminate and tungsten[3] and the moulded cathode mentioned by MacNair, Hannay, and Lynch[4]. The cathode to be described combines some of the features of the barium-aluminate cathodes and of the moulded cathodes in that it may consist of a homogeneous mass of active material as does the barium-aluminate cathode but has as basic material mixtures of alkaline-earth carbonates and iron nickel or cobalt powder with addition of suitable reducing agents. During manufacture the mixed powders are compacted at high pressures, usually into metallic retaining cylinders so that handling is simplified. During decomposition of the alkaline-earth carbonates to oxides and alkaline-earth metal, the nickel sinters into a strong metallic matrix. A diagrammatic cross-section of several possible forms of the cathode are shown in Figure 1, variations of the percentage of nickel powder may be used to vary the porosity of the nickel matrix depending on the requirements for the cathode.

Since the emissive surface of the cathodes studied most was probably barium or barium oxide on nickel, the cathodes have been called somewhat loosely Bariated-Nickel or *BN* cathodes.

2. *Materials*

The iron or nickel powder used is obtained commercially as carbonyl powder by decomposition of the metallic carbonyl. It is very pure and consists of small almost spherical particles of 1 to 5 microns in diameter. The cobalt powder is obtained by reduction from the oxide. The particles are less regular in shape but are about the same size. The packing under pressure of the cobalt powder is rather less than that for the nickel or iron powders for the same pressure.

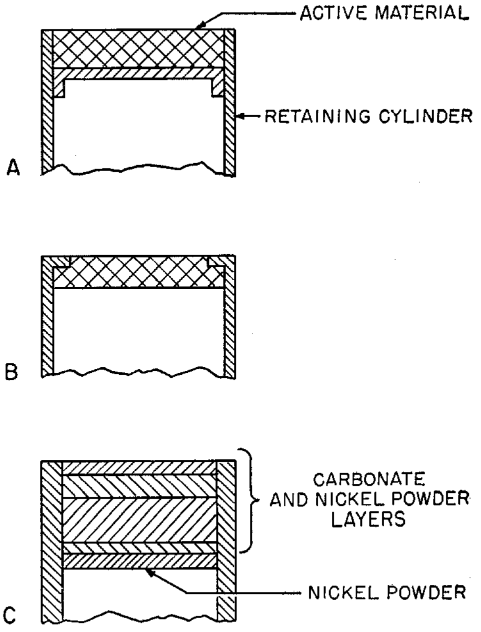


Figure 1—Mechanical construction of *BN* cathodes.

The metallic powders are washed, ball-milled with acetic acid, rewashed with distilled water, and vacuum stoved.

The alkaline earth carbonates (either barium carbonate, co-precipitated barium–strontium carbonate, or barium–strontium–calcium carbonates) are those used in the preparation of normal oxide-cathode coating. Various reducing agents may be employed: tungsten, tantalum, titanium tend to produce excess barium evaporation, whereas carbon reacts very slowly with prolonged evolution of gas. Zirconium, usually in the form of zirconium hydride (the commercial name for a solution of hydrogen in zirconium), has given best results.

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* Reprinted from *Le Vide*, volume 9, pages 302–309; November, 1954.

[1] H. J. Lemmens, M. J. Jansen, and R. Loosjes, "New Thermionic Cathode for Heavy Loads," *Philips Technical Review*, volume 11, pages 341–350; June, 1950.

[2] G. A. Espersen, "L-Cathode Structure," *Proceedings of the IRE*, volume 40, pages 284–289; March, 1952.

[3] R. Levi, "New Dispenser Type Thermionic Cathode," *Convention Record of the IRE 1953 National Convention*, Part 6—Electron Devices—Engineering Management, pages 40–41; 1953.

[4] D. MacNair, R. T. Lynch, and N. B. Hannay, "Molded Thermionic Cathodes," *Journal of Applied Physics*, volume 24, pages 1335–1336; October, 1953.

A large number of metals or alloys have been used for the retaining cylinder. With nickel or low-heat-conductive nickel alloys such as kovar, adhesion problems are a minimum. Molybdenum, used because of its low thermal emissivity, requires special treatment (that is, sintering nickel powder onto the part that will be in contact with pressed emission material) in order to ensure good adhesion between the cylinder and the nickel matrix.

The backing plate shown in Figure 1A is only used when heater–cathode insulation is a problem. It may consist of nickel or a nickel alloy, preferably not containing added reducing agents such as magnesium.

The metal cylinder and backing plate if used are hydrogen stoved before use.

3. Preparation of Cathodes

A mixture of 1 part by weight of reducing agent with 30 parts dry carbonate powder is made up. A powder containing 20–40 per cent of this mixture, the remainder being metal powder, is then made up and ball-milled in amyl acetate. The composition of the mixture is varied depending on the final cathode requirements. Large percentages of metal and high pressures give very robust cathodes with very little or no shrinkage on sintering the pressed mixture. Large percentages of carbonates give slightly enhanced emission but are less strong mechanically. Unless otherwise stated, results quoted for these cathodes will refer to the most extensively studied mixture containing roughly 30 per cent by weight co-precipitated double carbonate and 70 per cent nickel powder.

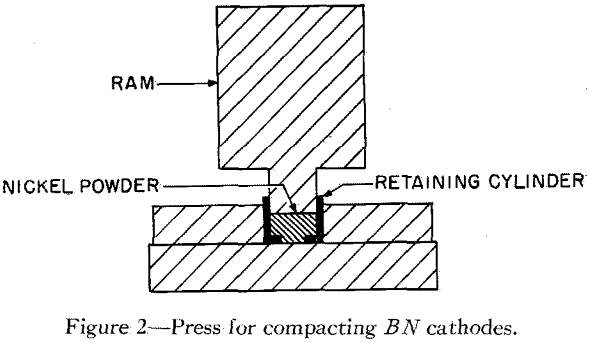

Figure 2—Press for compacting *BN* cathodes.

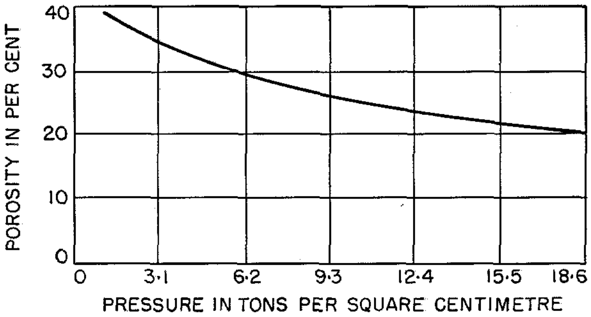

Figure 3—Porosity of pressed powders as a function of pressure.

The mixed, ball-milled powders are dried and compressed in the retaining cylinder as shown in Figure 2. The hardened steel blocks that form part of a hydraulic press have those faces that are in contact with the powder polished, both to prevent adhesion of the pressed powders and for ease of cleaning. The ring is also hardened and is used to prevent lateral expansion of the powder and retaining cylinder. This retaining cylinder may have a lip on its working face as shown in Figure 1B to improve the adhesion of the pressed powder. The backing plate if used is fitted on the nose of the hardened tool and pressed into place during the pressing operation. It is welded to the retaining cylinder after withdrawal from the press.

The pressure used to compact the powders may vary considerably with the composition of the mixed powders. A curve of applied pressure against porosity of the pressed mass of standard 30-per-cent-carbonate mixture is shown in Figure 3. For pressures greater than 50 tons per square inch the change of porosity and therefore change of linear dimensions on sintering is very small and pressures between 50 to 100 tons per square inch are used for the standard mixture and for most other mixtures—the pressures for mixtures containing large percentages of cobalt are, however, usually higher.

The cathode complete with retaining cylinder is ejected from the steel ring and is ready for mounting into a valve. If it is to be stored for a long time it is kept in a vacuum desiccator.

The pressed powders have a bright metallic surface, are easy to handle, and are sufficiently robust to allow a certain amount of surface machining, if required.

Although this description has only considered planar cathodes it is clear that initial shaping of the parts could produce cathodes of a large variety of shapes.

4. *Activation*

Decomposition of the carbonates and activation of these cathodes is very simple and follows very closely the schedule for standard oxide cathodes. The time required for pumping is very little greater than that required for an oxide cathode. In a typical instance the *BN* cathode can be outgassed in about ten minutes whereas any oxide cathode of similar dimensions would take two minutes. Emissions of 1 ampere per square centimetre are drawn while pumping.

After pumping the emission is normally stabilized by ageing the cathodes for 24 hours drawing 1 ampere per square centimetre. The cathode temperature is reduced from 1100 to 950 degrees centigrade during the first two hours of ageing and maintained at 950 degrees for the rest of the ageing time.

Since, for a plug of active material 1 millimetre thick, the mass of carbonate included is about 20 times the mass of carbonate on an oxide cathode of the same area, the amount of gas evolved during carbonate decomposition is considerable. In some cases it is therefore desirable to decompose the carbonates in vacuum before assembling the cathode in its final valve. Activated cathodes may be opened to air and stored in dry air for several days. On repumping the amount of gas evolved is very greatly reduced and reactivation follows the normal activation schedule.

After carbonate decomposition, the nickel matrix has sintered and the cathodes are mechanically very strong. The active material is easy to machine or polish and connections may be welded to it. Attempts to roll it into thin sheets have, however, not proved successful as the material becomes very brittle by the time it has been reduced to half its original thickness by rolling.

5. *Emission Properties of BN Cathodes*

In Figure 4 are summarized the pulsed and direct-current emission characteristics of *BN* cathodes as a function of true cathode temperature (that is, pyrometer temperature corrected for emissivity of the active material which, to within experimental error, is the same as nickel for nickel-powder-based cathodes). For comparison pulsed curves representative of good oxide cathodes and the published curves for Lemmens cathodes are shown.

The shaded area in Figure 4 represents the spread found in the pulsed emission results for fully activated *BN* cathodes. Also shown in Figure 4 is a line indicating the generally accepted practical values for the temperature of oxide cathodes for given continuous operating conditions. A similar line has been tentatively

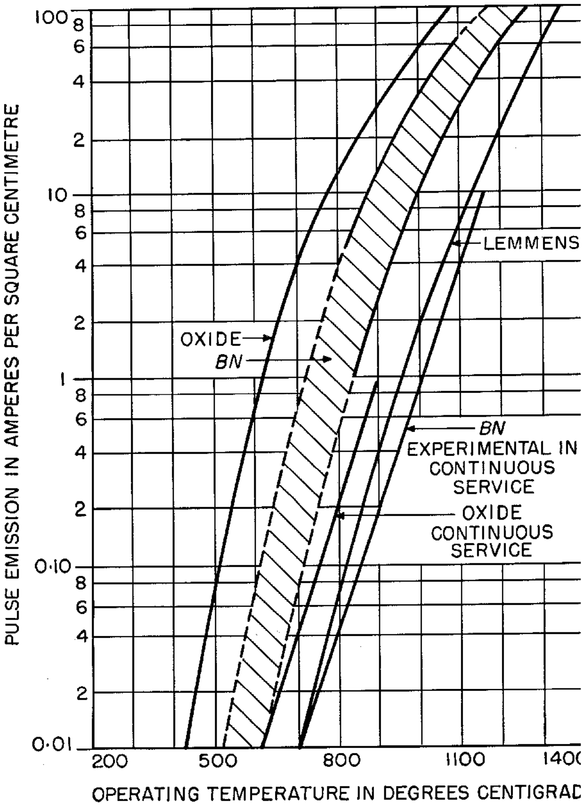

Figure 4—Emission as a function of temperature for oxide cathodes, *BN* cathodes, and Lemmens-type cathodes.

drawn for *BN* cathodes. The exact location of this line clearly depends on extensive life-testing. For the same emission density, *BN* cathodes require a temperature nearly 100 degrees centigrade hotter than the oxide cathodes. In compensation, the lives of *BN* cathodes are greatly prolonged. At 1 ampere per square centimetre and an operating temperature of 1000 degrees centigrade, lives of 5000 hours are obtained and at 1070 degrees centigrade and 3 amperes per square centimetre lives of 3000 hours are obtained.

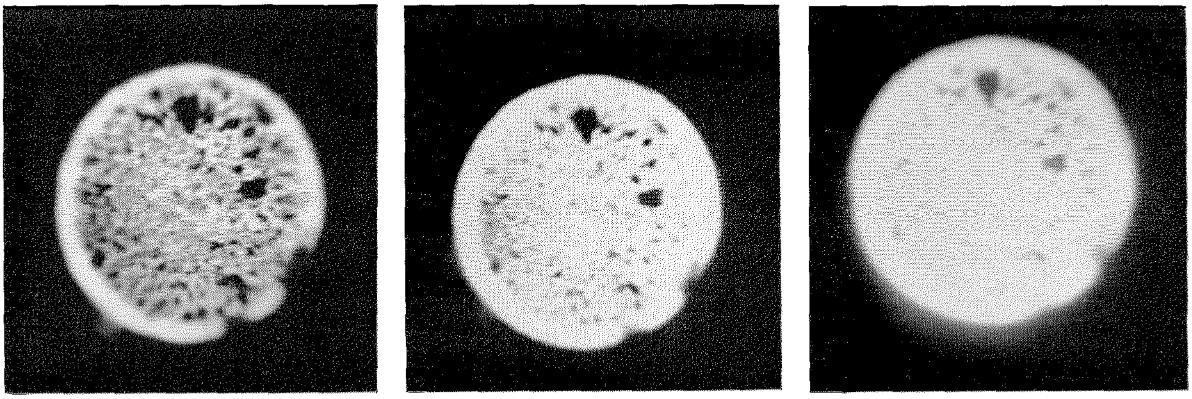
Figure 5—Images of *BN* cathode obtained by electron-microscope technique. Left, before activation. Center, during activation. Right, after activation.

As mentioned earlier, their recovery from oxygen contamination is remarkable and their resistance to poisoning in general is very high. A cathode operated in a pressure of air of about 10^{-4} millimetre of mercury gave an emission of 1 ampere per square centimetre for over 500 hours. *BN* cathodes have been extensively used in demountable vacuum systems with excellent results.

6. Emission Mechanism

There are two possible physical pictures of the emission properties of these cathodes: either the cathode is a normal oxide-cathode emitter emitting from small areas between the nickel powder but having a large reserve of free barium to give the required impurity centres for semi-conduction, or it is a genuine diffusion cathode with metallic barium or possibly barium oxide diffusing over the surface of the nickel giving a barium-on-nickel emitting surface. Experimental evidence seems to suggest that the latter is most probable.

Attempts to measure an interface resistance between nickel and oxide material has given no measurable resistance after several thousands of hours even though resistances of less than 5 ohms per square centimetre could be detected. In fact the resistance of the plug is so low that even at continuous emission densities of 3 amperes per square centimetre the cathode cooling due to electron evaporation is greater than the resistive heating and a drop of temperature of up to 20 degrees centigrade is observed on a standard life-test cathode. This in itself is not conclusive evidence as the large area of nickel-oxide surface contact and the volume of free barium would all tend to decrease the apparent interface resistance.

However a measurement of the velocity distribution of electrons at high current densities using the technique of Loosjes, Vink, and Jansen[5] showed no measurable ($<$ 5 volts) distribution of electron energies from these cathodes. Since according to Nergaard[6] this is a property of the emitting surface of an oxide cathode rather than of the oxide–metal interface, the semi-conductor hypothesis seems unlikely.

Furthermore examination of the emitting area of the cathode using a cathode-ray tube as an elementary electron microscope and forming an enlarged image of the cathode on the phosphor shows that the emitting areas of the cathode increase greatly during activation. Photographs of a cathode early in activation, partially activated and fully activated are shown in Figure 5 for a magnification of about 250. It is clear that diffusion of emitting material over the surface of the cathode is included in the mechanism of activation.

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[5] R. Loosjes, H. J. Vink, and C. G. J. Jansen, "Thermionic Emitters under Pulsed Operation," *Philips Technical Review*, volume 13, pages 337–345; June, 1952.

[6] L. S. Nergaard, "Studies of the Oxide Cathode," *RCA Review*, volume 13, pages 464–545; December, 1952.

7. *Work Function and Surface Coverage*

In an attempt to clarify the emission mechanism the work function of a *BN* cathode has been measured in 3 different ways.

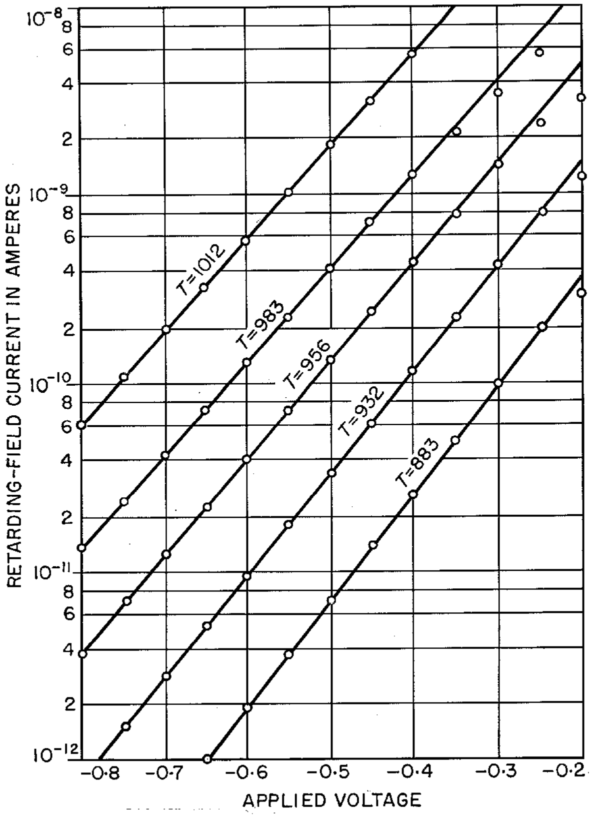

Figure 6—Retarding-field plots obtained from molybdenum cathode. *T* is the temperature in degrees Kelvin.

7.1 Retarding-Field Method

The method described by Benda[7] was used. In this method the *BN* cathode is made up into a diode facing a molybdenum-plate anode that is provided with a heater. In the activation of the *BN* cathode, free barium is deposited on the molybdenum anode, which is subsequently used as a cathode for the retarding-field measurements. It can be shown that

$$i_a = A_a\, F_a\, T_a^2 \exp\left[e\,(V - \phi_c)/kT_a\right].$$

Here the suffix *a* refers to the molybdenum "anode."

A_a = anode emission constant
F_a = anode area
T_a = absolute anode temperature
V = applied anode–cathode voltage
ϕ_c = arithmetic-mean work function of *BN* cathode.

Log i_a is plotted as a function of V for several values of T_a. The values i_{ao} corresponding to $V = 0$ are determined by extrapolation of the

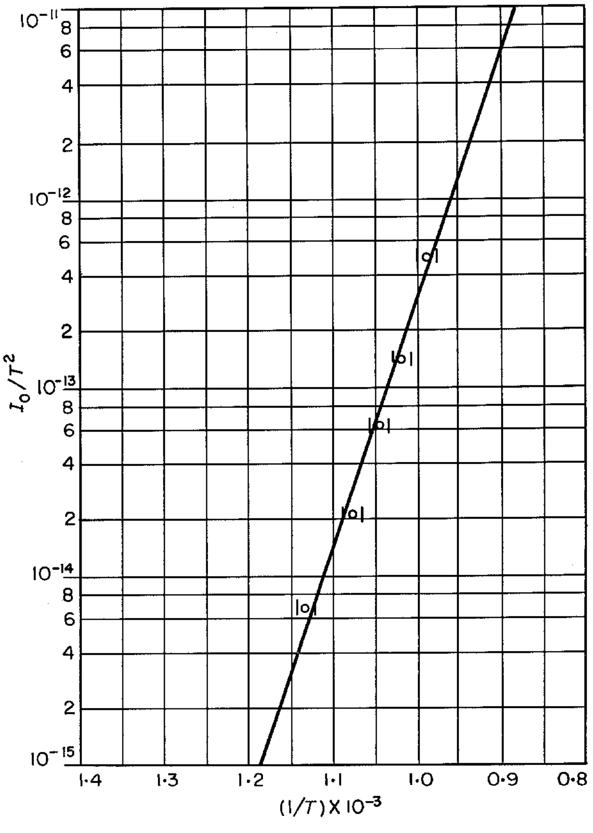

Figure 7—Richardson curves for *BN* cathode obtained by retarding-field measurements. *T* is the temperature in degrees Kelvin.

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[7] H. Benda, "Die Emissionskonstanten von Metall-Kapillar-Kathoden," *Frequencz*, volume 7, pages 226–232; August, 1953.

straight-line parts of the curves and ϕ_c is determined from a plot of log (i_{ao}/T_a^2) against $1/T_a$, as usual. The values of T_a are obtained from the slope of the original lines.

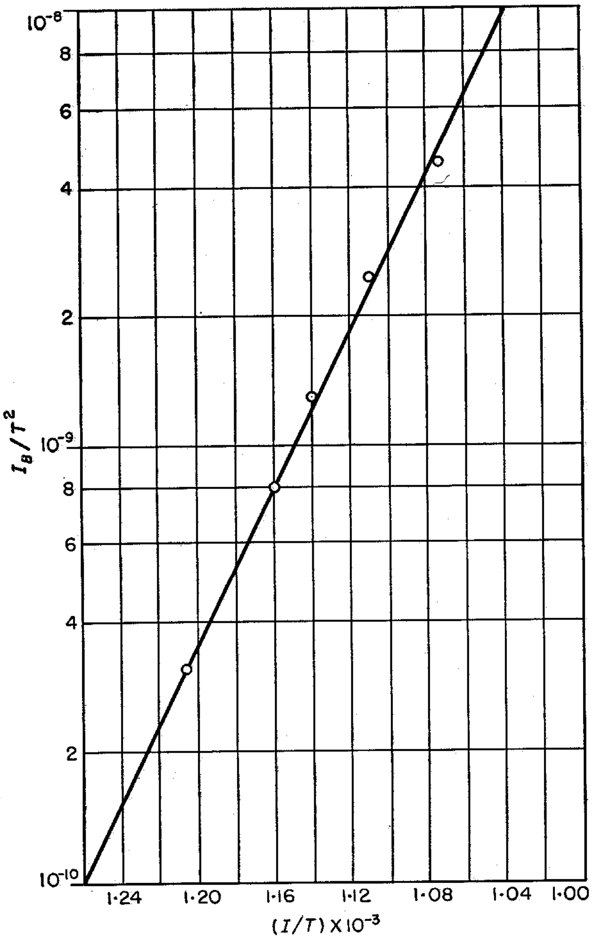

Figure 8—Richardson curve for *BN* cathode.

7.2 Richardson Plot

The data for this plot are taken using both pulse and direct-current operation. The saturation current is assumed to be the current at which the diode characteristic breaks away from a straight line on 2/3-power-law paper. It has been extablished that a Schottky plot gives a zero field current value lower than is obtained from the space-charge break and the temperature deduced from the Schottky line is very low. The zero field current deduced from a Schottky plot is thus unlikely to represent the true saturation emission.

7.3 Comparison with Charts of Loosjes and Jansen[8]

These charts assume that $A = 120$ amperes per square centimetre and that the cathode area is given. If the saturation emission (in our case, the results of the experiment above described) and the temperature are known, the charts yield a value of work function that depends on the temperature. If the work function is plotted as a function of T, the results should show whether there is a systematic variation of ϕ with T and should allow the zero-temperature value of ϕ to be deduced if a straight-line law is obeyed.

8. Results

Typical experimental results for the three methods are shown in Figures 6 through 9.

The mean work function measured by the method of section 7.1 yields a value of $\phi_c = 2{\cdot}8$ electron volts. Method 7.2, which as is well known, yields a value for the minimum work function of a patchy surface gives $\phi_{\min} = 1{\cdot}78$ electron volts. Method 7.3 yields a value about 1·91 electron volts which, when plotted against T, yields a value of $\phi_o = 1{\cdot}79$ electron volts, is in good agreement with the Richardson value. The temperature coefficient of work function is $1{\cdot}36 \times 10^{-4}$ electron volt per degree Kelvin.

Now, let us assume that the cathode can be represented as a patchy surface having two distinct values of work function. One of these has the value 1·78 electron volts and the other is that of nickel, 5·0 electron volts. Let the proportion at the low work function be a. Then,

$$\phi_c = a\,\phi_{\min} + (1 - a)\,\phi_{\text{nickel}} \qquad (2)$$

And, inserting values, $a = 0{\cdot}69$. Thus, the surface coverage is in the region of 70 per cent.

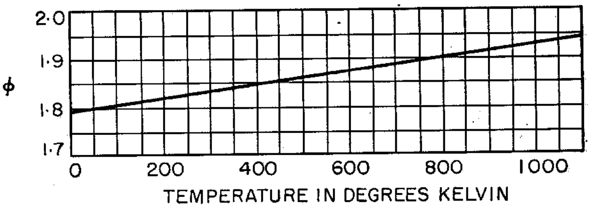

Figure 9—Work function for *BN* cathode plotted against temperature. From the curves of Loosjes and Jansen.

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[8] C. G. J. Jansen and R. Loosjes, "Graphs for Rapid Calculation of Work Function of Thermionic Emission," *Philips Research Reports*, volume 8, pages 81–90; April, 1953.

We therefore conclude that the *BN* cathode can satisfactorily be represented as a barium-on-nickel diffusion emitter in which the surface coverage is about 70 per cent, which is in approximate agreement with the electron-microscope photographs, and in which the temperature coefficient of work function is about 10^{-4} electron volt per degree Kelvin.

9. Conclusions

In addition to the use of these cathodes in testing diodes, they have been used in reflex klystrons and television cathode-ray tubes and have given good service in both applications.

It is considered that the *BN* cathode should find a very wide field of application, especially in valves for ultra-high frequencies.

Two New Equations for the Design of Filters

By MILTON DISHAL

Correction for Volume 30, Pages 324–337; December, 1953

IN THE FIRST LINE of Step 8 on page 329, change "left" to "right".

In the numerator of (6), change the plus signs of the second, fourth, sixth, et cetera terms to minus signs.

In the first line of Step 11 on page 330, change "right" to "left" and in the second line replace the first 10 words with "all terms will be positive."

The above changes in no way affect the synthesized network since the actual synthesis is accomplished by using (7) (and (8)). However, it is true that the right-half-plane zeros of Step 7 produce (7) and the left-half-plane zeros of Step 7 produce (8).

There is the following additional point involved: it is sometimes necessary or desirable to know the *terminated* voltage (current) reflection coefficient or the terminated input impedance (admittance) of a particular pair of terminals.

As indicated in Step 10, (7) gives the short-circuit (open-circuit) input immittance of *one pair* of network terminals. The terminated voltage (current) reflection coefficient (complex) of *this same pair of terminals* is given by the ratio of the numerator of (6) (as corrected) to the numerator of (5). The point being stressed is that there must be alternating signs in the numerator of the terminated reflection coefficient (complex) for that pair of terminals that has (7) as its short-circuit (open-circuit) input immittance. Similarly, in agreement with the corrections to Step 11, there must and will be all positive signs in the numerator of the terminated reflection coefficient (complex) for that pair of terminals that has (8) as its short-circuit (open-circuit) input immittance.

The above correction could have been accomplished simply by interchanging (7) and (8). This procedure has not been followed because it is important for practical reasons to use the present (7) for the major part of the synthesis. This is true because (as can be seen) the denominator of (8) is formed by taking *differences*, and significant figures are usually lost in the process.

The author is indebted to Mr. E. Green of Marconi's Wireless Telegraph Company, Limited, for discussions (via mail) that clarified the above points.

Novel Gas-Gap Speech Switching Valve *

By A. H. BECK, T. M. JACKSON, and J. LYTOLLIS

Standard Telecommunication Laboratories, Limited; London, England

A NEW gas-discharge circuit element is described. It has the following properties; low resistance (about 100 ohms) when struck, the resistance being constant for all frequencies up to at least 50 kilocycles per second, the reactive component of the impedance is negligible, and the noise voltages are at least 60 decibels below signal levels. The development of such devices considerably simplifies major problems in electronic telephone exchange design.

. . .

The all-electronic telephone exchange has been the subject of much research in the last decade. One of the simpler ways of making such an exchange seems to be the direct replacement of the crossbar switch by electronic devices, each junction of the mechanical switch being replaced by a suitable electronic device. The more obvious requirements of such a device are:—

A. That it can be switched rapidly from a high (practically infinite) alternating-current impedance to a low (practically zero) alternating-current impedance.

B. That it shall transmit voice frequencies at milliwatt power levels without either frequency or amplitude distortion.

C. That it shall not introduce electrical noise.

D. The device should form part of the line circuit and the switching operation should be performed by pulses in the line circuit.

A cold-cathode gas-filled valve is an obvious device to investigate for the purpose outlined because the glow discharge can be initiated and extinguished directly by pulses and there is a definite break between the conducting and non-conducting state within a time of the order of 1 millisecond. There are other advantages in using a cold-cathode gas-filled valve. When in a non-conducting state the valve consumes no power; rapid fault diagnosis is possible because the glow is easily visible; and such valves have long lives.

The main difficulty to be overcome is that of providing a sufficiently low resistance path for reasonable direct currents in the conducting state. Bell Telephone Laboratories[1] have produced a diode gas-gap using a hollow-cathode discharge, which gives a small power gain in the frequency range of 50 to 3000 cycles per second and operates with direct currents in the range 10 to 20 milliamperes. It is not known whether this valve can be switched and used directly in the line. This article describes an electrode arrangement for gas-filled valves that gives a noise-free, low-resistance path for direct- and alternating-current signals, the upper frequency for the alternating-current signals being at least 50 kilocycles per second. This device satisfies the other requirements stated in this introduction. Several different electrode structures are possible and two practical valves using the arrangement are described.

1. *Description and Mode of Operation*

The electrode arrangement consists essentially of a cathode with two electrodes acting as joint anodes equally spaced from the cathode surface. These anodes can be of any geometrical shape provided a normal projection from them in the direction of the cathode will meet the cathode surface. There is a limit to the anode–cathode separation, depending on the nature and pressure of the gas filling. Figure 1 shows various electrode arrangements that have been found to work satisfactorily. Figures 2 and 3 show the variation of anode-gap resistance with cathode current and frequency respectively. The gap

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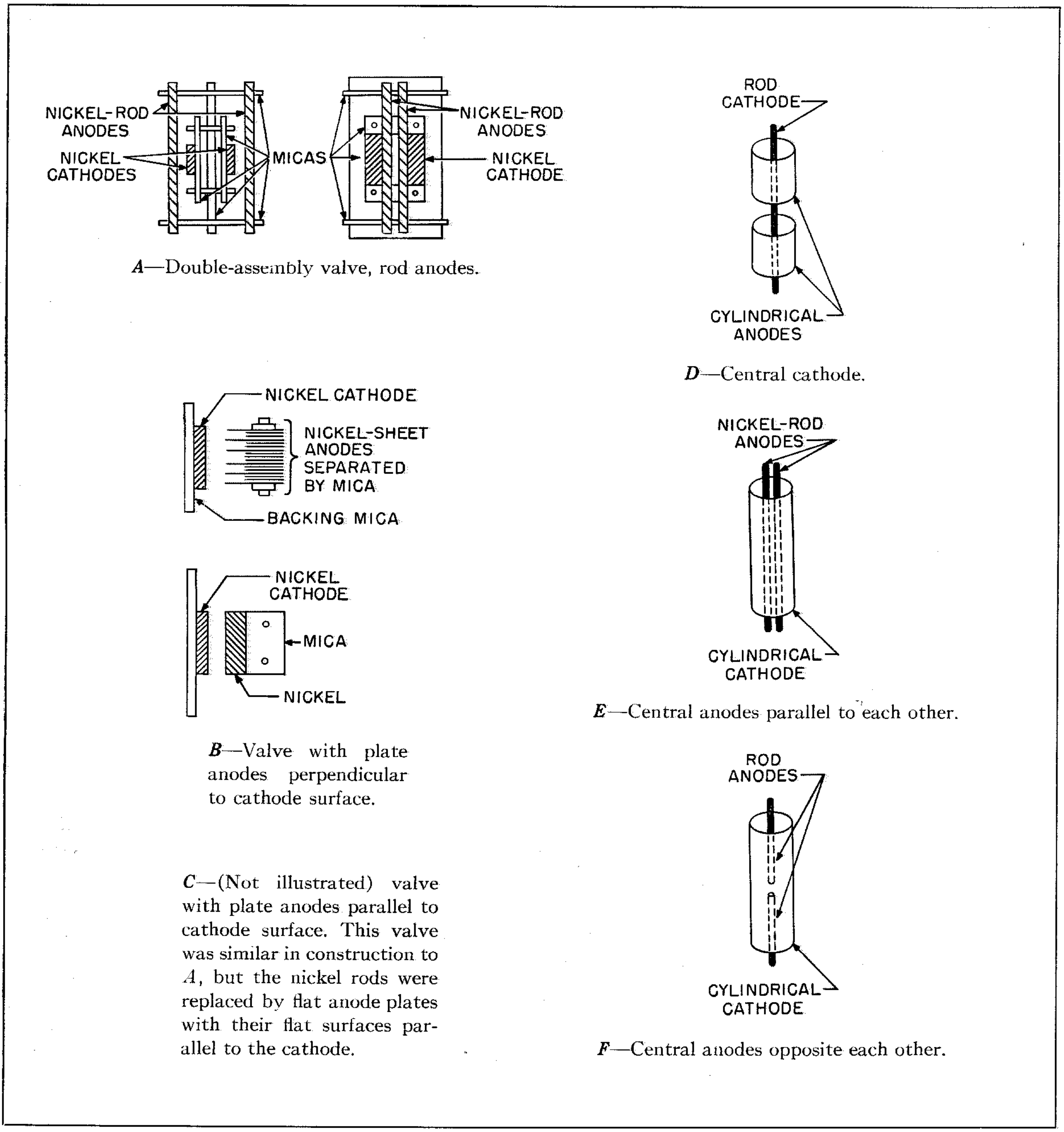
* Reprinted from *Electronic Engineering*, volume 27, pages 7–12; January, 1955.

[1] M. A. Townsend and W. A. Depp, "Cold Cathode Tubes for Transmission of Audio Frequency Signals," *Bell System Technical Journal*, volume 32, pages 1371–1391; November, 1953.

resistance can be measured directly using an alternating-current bridge. It can be seen that the gap resistance is substantially independent of frequency but varies inversely with the total current through the valve. The apparent increase of resistance at low frequencies is because the impedance of blocking capacitors used in the measuring circuit becomes appreciable at these frequencies. Below a current of 10 milliamperes there is a rapid increase of gap resistance but above this current, the variation of resistance with current is gradual. The gap resistance is substantially non-reactive, there being no frequency-dependent inductive component present such as is normally found in diode electrode arrangements, It is also noise-free.

Figure 4 shows the direct-current characteristics of a typical valve using the electrode



A—Double-assembly valve, rod anodes.

D—Central cathode.

B—Valve with plate anodes perpendicular to cathode surface.

E—Central anodes parallel to each other.

C—(Not illustrated) valve with plate anodes parallel to cathode surface. This valve was similar in construction to *A*, but the nickel rods were replaced by flat anode plates with their flat surfaces parallel to the cathode.

F—Central anodes opposite each other.

Figure 1—Various structures for double-anode valve.

arrangement already described and Figure 5 shows the circuit used to measure the characteristic. The anodes were controlled by separate high-voltage supplies and these supplies were first adjusted so that the currents in the two anode circuits were equal. The voltage of each supply was then adjusted so that one anode source was raised in voltage by steps of 5 volts, while the other anode source was lowered in steps of 5 volts. The currents in each anode circuit were measured for each 5-volt variation. and the voltage difference between the anodes was also measured. The characteristic shows that about the point where the anode currents are balanced, there is a region where a small change in voltage on one anode produces large current variations in the two anode circuits. It is also obvious that as the current in one anode circuit increases, there is a corresponding decrease in the other anode circuit of approximately the same amount, so that the cathode current remains constant. The linearity of the graph in the working region shows that distortion will be negligible. The graph also shows that this electrode arrangement has the important property that a surge of voltage in one anode circuit will not extinguish the discharge so that the valve ceases to operate. Such a surge will merely cause one anode to stop taking current, the whole of the current through the valve being taken by the other anode until the surge has passed. This is a useful characteristic for telephone line operation.

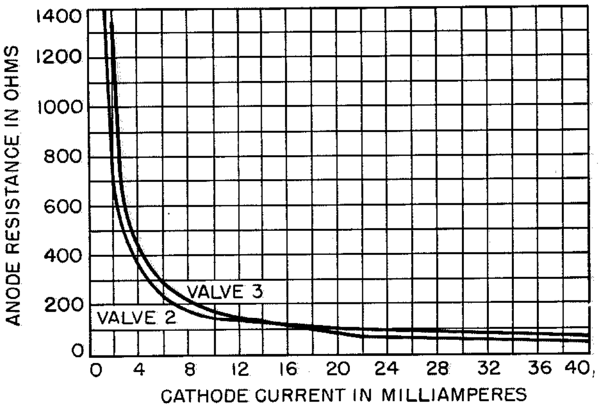


Figure 2—Variation of resistance with cathode current. Valve *2*—0·5-millimetre gap between anodes. 30 millimetres of mercury pressure of neon–argon gas mixture. Valve *3*—0·5-centimetre gap between anodes. 30 millimetres of mercury pressure of neon–argon gas mixture.

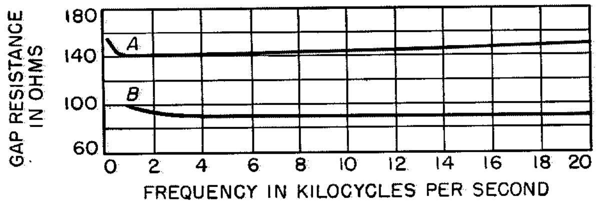


Figure 3—Variation of gap resistance with frequency. Valve *2*—0·5-centimetre anode gap. Neon-argon gas mixture at a pressure of 30 millimetres of mercury. Curve *A* is for 10 milliamperes and curve *B* 20 milliamperes total cathode current.

Valves incorporating the double-anode electrode arrangement can be used for speech transmission in a circuit similar to that shown in Figure 6. The valves are acting effectively as switches in the line. In a 10-kilohm-impedance

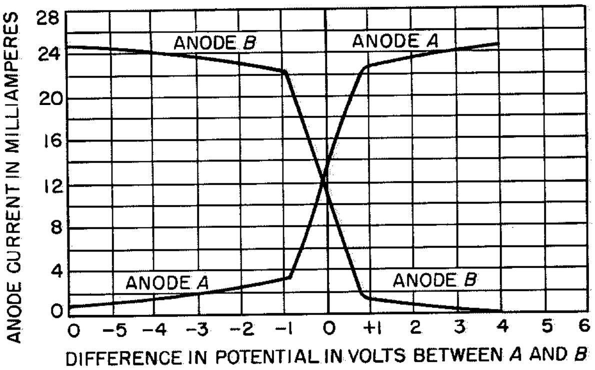


Figure 4—Anode characteristic of double-anode valve. Valve *3*—0·5-centimetre anode gap. Neon–argon at 30 millimetres of mercury pressure.

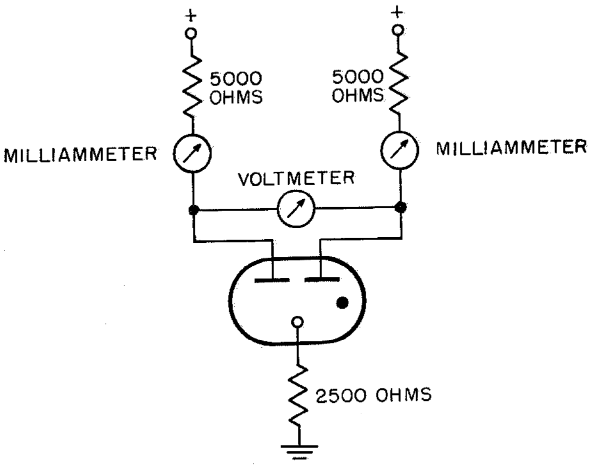


Figure 5—Test circuit for direct-current characteristic.

circuit the total loss due to a pair of valves acting as one stage is less than 0·2 decibel.

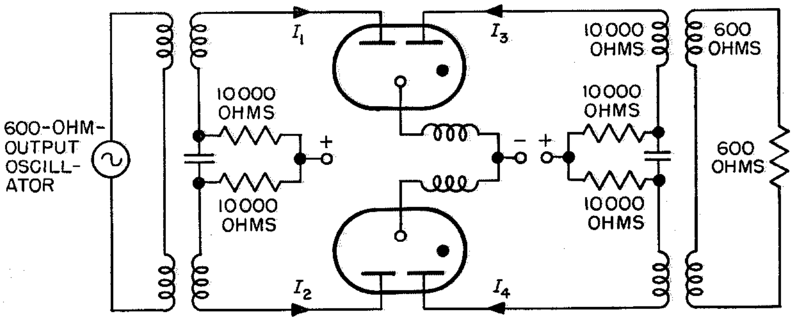


Figure 6—Use of the double-anode valve as a switch directly in the telephone line. I_1, I_2, I_3, I_4 = 10 milliamperes.

To enable valves using the electrode arrangement to be switched directly using pulse voltages in the line, it is necessary to have stable, reproducible, electrical characteristics. For instance, the maintaining voltage V_m of all valves should be within the limits of ± 5 volts if V_m is of the order of 100 volts and the breakdown voltages V_b should be within the limits of ± 10 volts if V_b is of the order of 250 volts. It is also necessary to have a large difference between the maintaining voltage and the breakdown voltage of the glow discharge. V_m is controlled by the nature of the gas filling, current density, cathode material, and state of the cathode surface. V_b is also controlled by these factors, but in addition it is a function of the product of gas pressure and electrode separation.

During study of the geometrical design factors necessary to ensure a low-resistance gap between the joint anodes, experiments were undertaken to relate gas pressure and electrode separation to gap resistance. This was done by taking a valve with an electrode separation of fixed value and measuring the resistance between the anodes when the valve was still on the pump bench so that the gas pressure could be varied at will. The current in the discharge was kept constant at 20 milliamperes. The results for helium are shown in Figure 7. It shows that there is a sudden sharp increase in resistance when a given pressure is reached for a particular electrode separation. Argon gave similar results. Figure 8 shows a similar graph for neon–argon–hydrogen gas mixture and here, after the sudden increase to a high value there is a rapid decrease to a fairly low resistance once more. It is interesting that there is normally a blue-coloured anode glow on the anode surface when there is a glow discharge in neon–argon–hydrogen gas mixture and yet this glow is not present at low pressures when a low anode-to-anode resistance is observed, but appears suddenly when the sharp increase in anode-to-anode resistance takes place. This argues the sudden formation of a space-charge sheath on the anode, because anode glows are formed when a space charge exists in the anode region. Probe measurements confirmed that space-charge sheaths do form on the anodes at the pressure at which the gap resistance suddenly increases to high values.

The above results show that the anode–cathode gap values cannot be increased indefinitely in order to secure a high V_b because there is a critical distance at a given gas pressure beyond

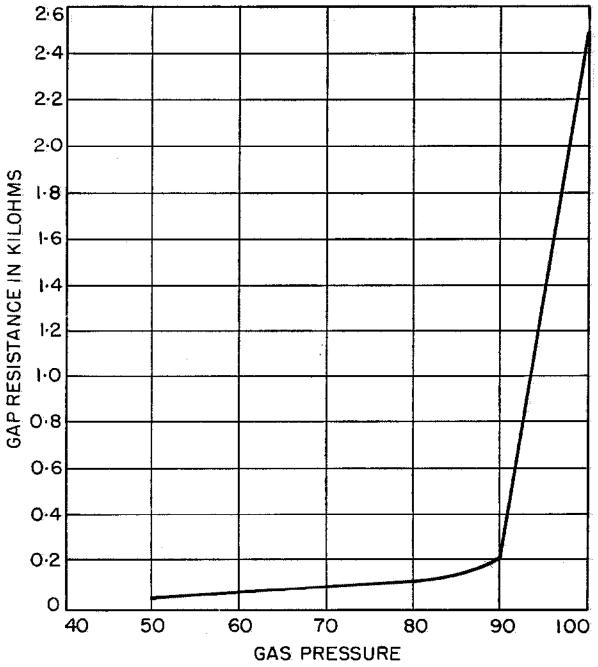


Figure 7—Variation of anode-to-anode resistance with gas pressure (helium). Gas pressure scale is in millimeters of mercury. Anode-to-cathode spacing was 2 millimeters.

which a low gap resistance cannot be obtained. However, it was found that this distance was sufficiently great for a large $V_b - V_m$ difference to be obtained while still retaining a low gap resistance if helium was used as a gas filling and the pressure was suitably chosen.

Another requirement when considering pulse breakdown is that the gap shall always break down when a pulse of given amplitude is applied to one or both of the electrodes.

There are two physical phenomena that introduce a time lag between application of a sufficient voltage on the anode to produce breakdown in a gas-discharge system and the stable establishment of the discharge. These are called the statistical delay time and the formative delay time.

The statistical delay time is dependent upon the presence of an electron or ion in the gap between the electrodes at a particular instant of time. The ion can be accelerated towards either electrode and produces sufficient ionization in the gas to cause breakdown between the electrodes. Ions are normally present in any gas atmosphere because of ionization by cosmic radiation, but the number of ions present per unit volume in a gas-gap can be greatly increased by irradiating the gas with X-rays or ultra-violet light or by photo-electric effects at the electrodes due to either of these agents or visible light. Thus, by suitable priming, the statistical delay time can be eliminated.

The formative delay time depends upon the time taken for the ion or electron that initiates the discharge to produce sufficient ionization to cause breakdown and produce a self-sustaining discharge. This depends upon parameters governed by the pressure and nature of the gas and geometrical dimensions that are normally constant for a given gas-gap, but it is also governed by the voltage applied between the electrodes. Thus, formative delay time cannot be eliminated, but is reduced by increasing the voltage applied between the electrodes to initiate breakdown above the value that will just cause breakdown if applied for a very long time.

Formative delay time necessitates the application of pulse voltages with amplitudes greater than the normal direct-current breakdown voltage of a gas-gap in circuits where pulse techniques are used to initiate the discharge. The difference in voltage between the pulse amplitude required to produce breakdown and the direct-current breakdown voltage is called the "over-voltage". Early measurements upon speech-gap tubes showed that the over-voltage required to produce breakdown depended upon:—

A. History of the cathode (that is, whether it has been conducting or left idle).

B. Pulse width.

C. Light falling on the electrodes; photo-electric effect.

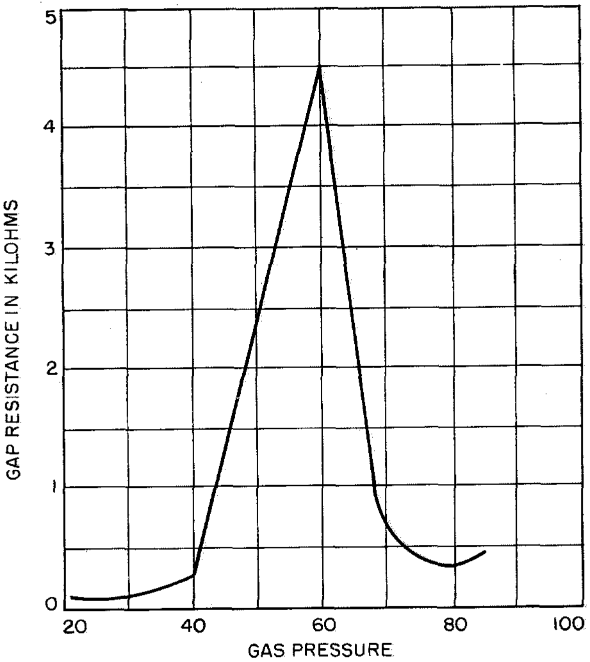


Figure 8—Variation of anode-to-anode resistance with gas pressure (neon-argon-hydrogen in the ratio 92:1:7). Gas pressure scale is in millimeters of mercury. Anode-to-cathode spacing was 2 millimeters.

There are several ways in which the photo-electric priming can be produced. An internal priming source can be incorporated in the valve, or an external agent such as an ultra-violet light source can be arranged to prime a bank of tubes. Another method of priming is to introduce a radioactive substance into the valve, so that ionization is produced by radioactivity. The availability of tritium, a weakly radioactive substance with a half-life of about 12 years and radiation energy of < 10 000 electron volts that can be used with safety within a glass envelope has greatly increased the attraction of this latter form of priming. The advantage of the method is that no electrical power is consumed to produce the priming.

2. *Theory of Low-Resistance Anode-to-Anode Gap*

Probe[2] and other[3] measurements recorded in original published papers have shown that the negative glow is a plasma region, there being high equal concentrations of positive and negative ions here. Concentrations of 10^9 ions per cubic centimetre have been calculated from measurements made, and Emeleus[4] quotes a figure as high as 10^{11} ions per cubic centimetre. The conductivity of the space between the anodes operating under the conditions outlined in this report is therefore high; a low resistance might be expected from this consideration. On the other hand, it is difficult to see why a change of one volt at either anode should cause the anode concerned to take all the current formerly being received by the other anode, which is normally only about 1 millimetre away.

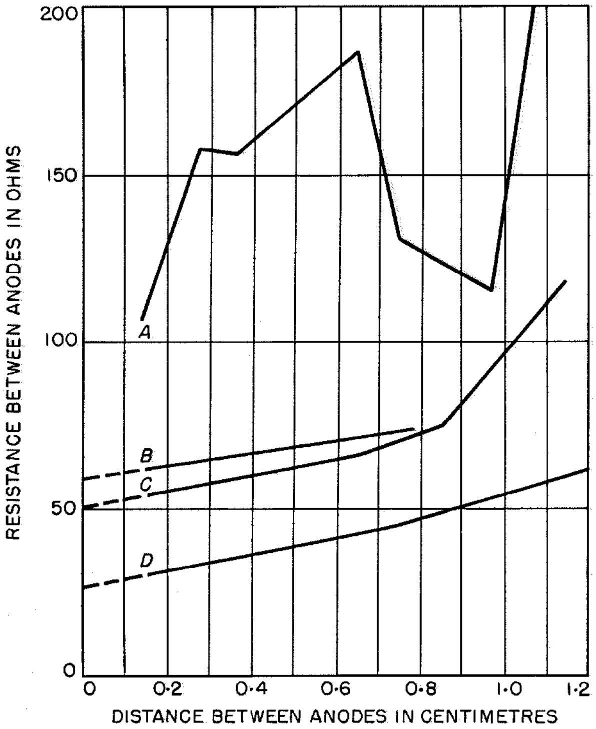


Figure 9—Variation of resistance with gap length. *A*—15 milliamperes per anode. *B*—20 milliamperes per anode; slope = 2·34 ohms per millimetre. *C*—20 milliamperes per anode; slope = 3·0 ohms per millimetre. *D*—30 milliamperes per anode; slope = 2·2 ohms per millimetre.

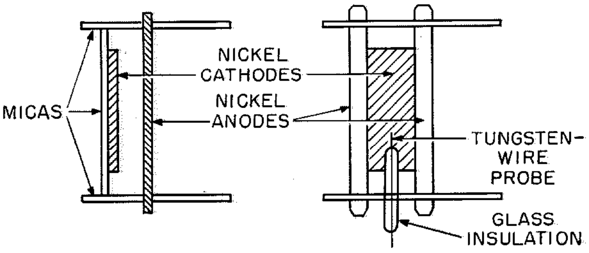


Figure 10—Double-anode valve with probe between anodes.

Figure 9 shows graphs of gap resistance against gap length for an experimental valve containing one fixed and one mobile anode. The electrode assembly was otherwise similar to that of normal valves except that a larger-area cathode had to be used and therefore high total currents were required to cover the whole cathode surface with a glow discharge. At a current of 15 milliamperes per anode an irregular curve was obtained. This was because there was a relatively dark patch on the cathode surface between the anodes; when the moveable anode was over this area the resistance between the anodes was high, but as soon as the anode moved over a part of the cathode where the negative glow was normal, the resistance of the gap decreased. At the maximum spacing the resistance increased suddenly because this was the edge of the negative glow region and the mobile anode was moving out of contact with the plasma. The graphs for higher currents show that there is a steady, small increase of gap resistance with gap width, this increase being only a few ohms per millimetre. The slopes of the graphs at the two higher currents are sensibly the same.

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[2] K. T. Compton, L. A. Turner, and W. H. McCurdy, "Theory and Experiments Relating to the Striated Glow Discharge in Mercury Vapor," *Physical Review*, volume 14, pages 597–615; December, 1924.

[3] B. Van der Pol, "Method of Measuring, Without Electrodes, the Conductivity of Various Points along a Glow Discharge and in Flames," *Philosophical Magazine*, volume 38, pages 352–364; September, 1919.

[4] K. G. Emeleus, "Conduction of Electricity Through Gases," Third Edition, Methuen and Company, Limited, London, England; 1951: see page 61.

The conclusions drawn from the results are that a field theory of the switching action cannot be substantiated, and that a theory on the conductivity of the plasma can only be advanced if some boundary condition at the anodes is postulated to account for the large residual resistance that is obtained for zero length when the graphs are extrapolated.

Probe measurements were next made to elucidate what happens to the plasma potential when one anode is raised and then lowered in potential so that it draws maximum current and then minimum current. A valve similar to that represented in Figure 10 was used and the measuring circuit is shown in Figure 11. The anodes were connected directly to the variable high-voltage sources and the probe potential was varied by means of a potentiometer supplied from a third high-voltage. The currents in the anode circuits were measured by low-resistance milliammeters and the probe current was measured by an accurate multi-range microammeter. The probe potential relative to earth was measured very accurately with a standard potentiometer and the same instrument was used to measure the anode voltages accurately. Since there was practically no resistance in the anode circuits, the shunt current did not affect the anode voltage when the potentiometer was connected between anode and earth. The probe potential measured by the potentiometer must be corrected for the potential drop across the microammeter.

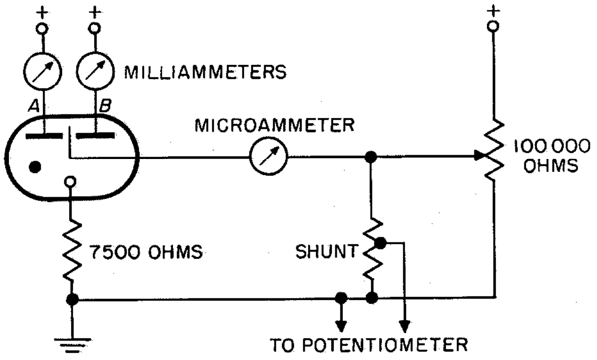

Figure 11—Circuit used for probe measurements.

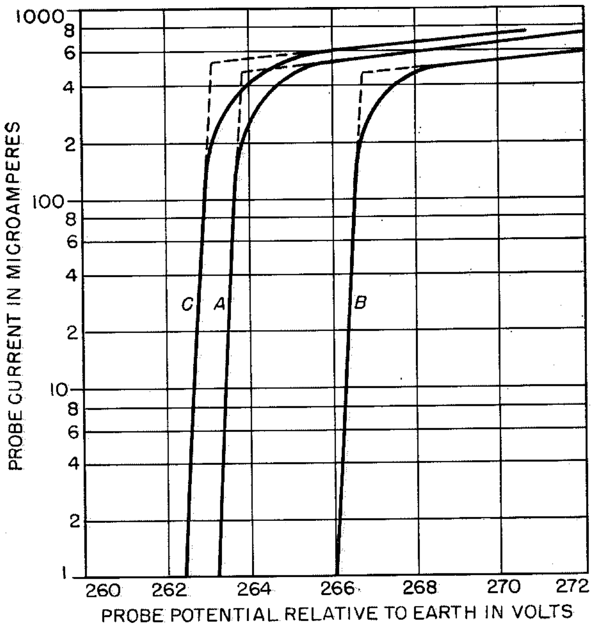

Figure 12—Probe characteristics for balanced and unbalanced states of anode currents in a double-anode valve.

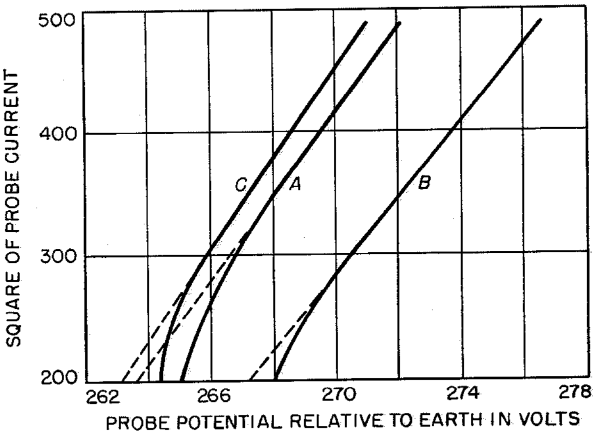

Figure 13—Probe characteristics for balanced and unbalanced states of anode currents in a double-anode valve.

The probe characteristics and anode voltages were measured *A*) with the anode currents balanced at 10 milliamperes, *B*) when anode *A* was raised in potential until anode *A* drew 20 milliamperes current and anode *B* drew zero current, *C*) when *A* was lowered in potential until *A* drew zero current and *B* drew 20 milliamperes current. Figure 12 shows the probe characteristics for these three states and Figure 13 shows the plots of the i^2/V characteristics for the same states for electron currents above the knee part of the probe characteristics. Langmuir probe theory requires that the i^2/V characteristic should be linear except near the space potential, and from the slope of the line the number of electrons per cubic centimetre of the plasma can be calculated.

The results were:—

A. Potential of anode *A* = 263·20 volts
Potential of anode *B* = 263·12 volts
Space potential (from graph) = 263·82 volts

B. Potential of anode *A* = 266·90 volts
Potential of anode *B* = 263·24 volts
Space potential (from graph) = 266·80 volts

C. Potential of anode *A* = 258·44 volts
Potential of anode *B* = 263·04 volts
Space potential (from graph) = 263·15 volts

Electron temperature in degrees Kelvin (from slope of graphs in Figure 12) = 1144

Number of electrons per cubic centimetre (from slope of graphs in Figure 13) = $3 \cdot 7 \times 10^9$

It is seen that the space potential appears to be positive with respect to the anode potential in the balanced state. Contact-potential difference between probe and anodes is concluded to be the cause of this.

The results now make clear the switching action of the double-anode valve. As one anode is raised in potential, the plasma potential follows the potential of the highest electrode with which it makes contact and leaves the other anode at a potential which is now negative with respect to the plasma potential; this latter anode takes a smaller electron current. The other anode takes a greater electron current because the high-voltage supply determines what the cathode current shall be; the current through the valve remains approximately constant. If the plasma potential rises to a voltage such that the lower-potential anode is several volts negative with respect to the plasma potential, then the current to this anode falls to zero; the other anode takes all the current passing through the valve. If one anode is lowered in potential, the plasma potential remains close to the potential of the other anode since it still follows the potential of the highest electrode with which it is in contact. The anode which is being lowered in potential therefore draws less and less electron current as it goes negative with respect to the plasma potential until the cut-off point is reached. The low electron temperature of the space around the anodes shows that the electrons have low energies and therefore a small negative difference of potential between the anode and the plasma will be sufficient to repel the electrons and reduce the electron flow to the anode.

If the anodes are not in intimate contact with the plasma then the above mechanism cannot work. This explains why a space-charge sheath on the anode causes a high anode-gap resistance. It also explains why a patchy glow produces a high gap resistance, because the anode over the part of the cathode where the negative glow is absent or of low intensity will have little influence on plasma potential.

3. *Practical Valves*

3.1 Multiple-Gap Valve

For convenience in circuit design, it is desirable to have a valve containing a number of gaps; a valve containing ten gaps is particularly suitable for telephone work. To meet this demand the multiple-gap valve shown in Figures 14 and

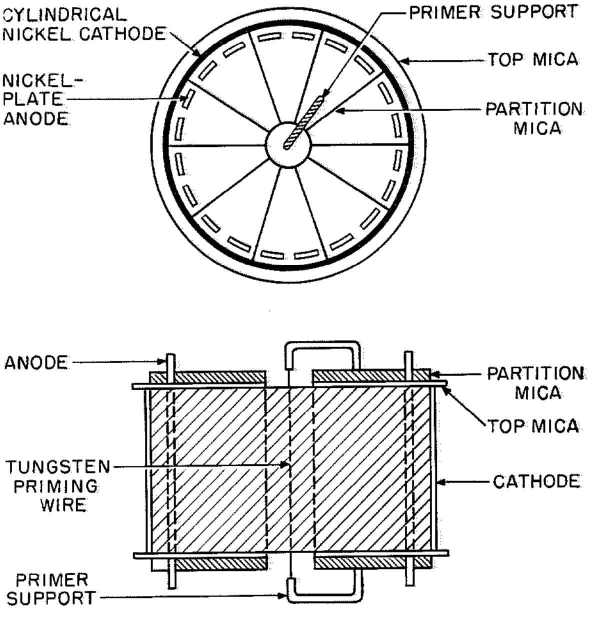

Figure 14—Multiple-gap valve.

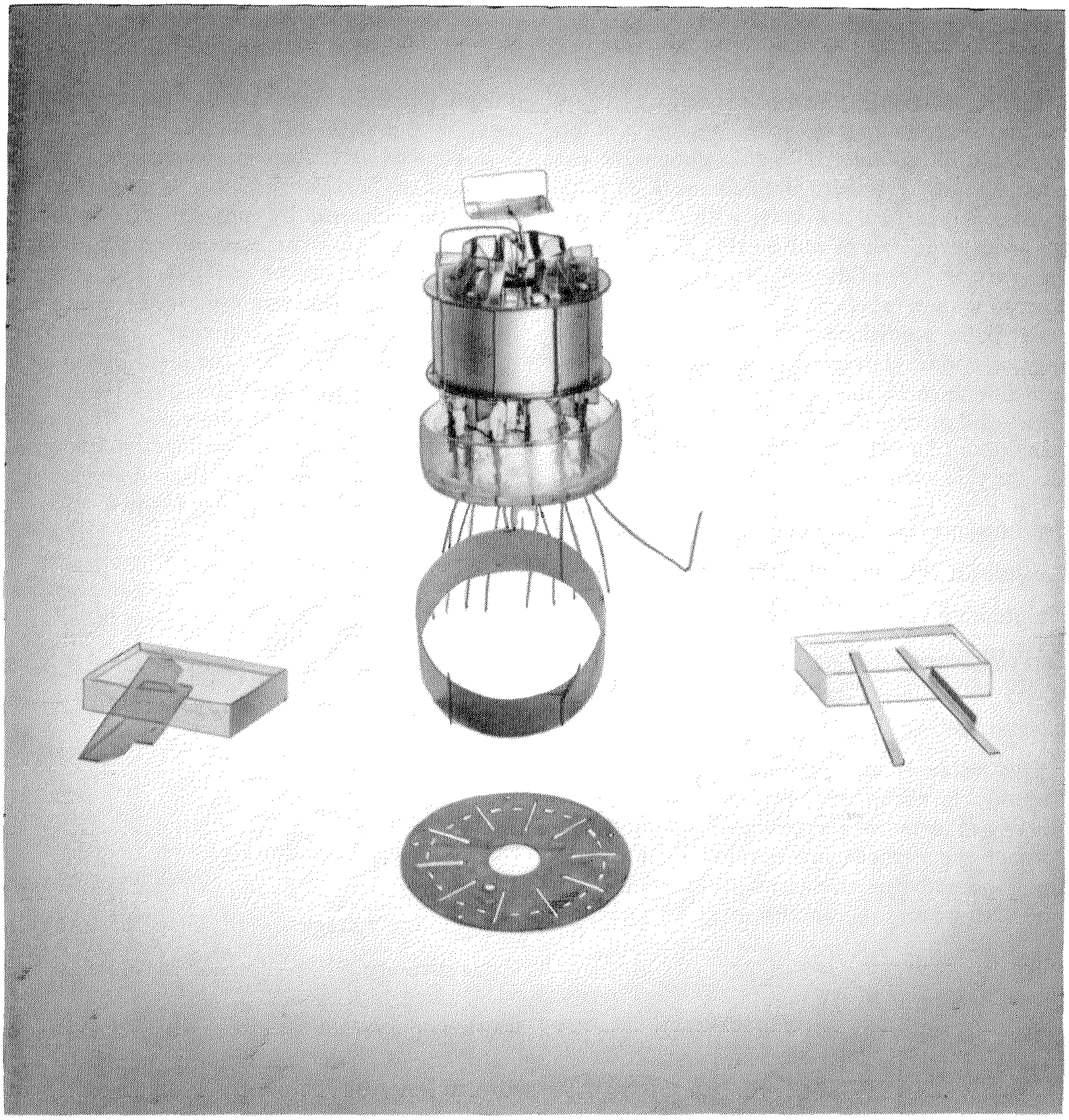

Figure 15—Multi-gap valve and component parts.

15 was devised. The assembly consists of a cylindrical common cathode made of nickel. The surface is divided into ten equal areas by vertical mica partitions that radiate from the axis of the cylinder to make a flush contact with the cathode surface. The partition micas are located in slots cut in the top and bottom micas that close the ends of the cathode cylinder. The anodes, two in each sector, are mounted at a distance of 2 millimetres from the cathode surface and are located in the end micas. At the centre of the cathode cylinder a vertical tungsten wire is mounted along the axis to serve as a photoelectric primer to all gaps. This priming device was for experimental purposes only, because although it is a simple means of priming, its power consumption is too high for practical purposes.

A number of these multiple-gap valves have been made and they have been found to operate satisfactorily in switching circuits. Great care is necessary to keep the cathode spacing constant at 2 millimetres in all ten gaps, because a difference of 0·25 millimetre in this dimension can cause a breakdown voltage difference of 20 volts. V_m is constant to a few volts within one valve, but over a number of valves this characteristic can vary considerably unless care is taken in processing and ageing. The valves work well under pulse conditions with pulse widths of 100 microseconds and overvoltages of 40 volts. There is a difference in the direct-current breakdown before and after a glow discharge, this difference being about 15 volts. Typical values of parameters for this valve are:—

Breakdown voltage before glowing = 330 volts

Breakdown voltage after glowing = 315 volts

Maintaining voltage = 165 volts

Gap resistance = 100 ohms

3.2 Single-Gap Valve

This valve is illustrated by Figure 16. The assembly is cheap and easy to produce. The pinch seal acts as a convenient mount. Both sides of the cathode are covered by the glow discharge and this eliminates the need for any micas. However, the anodes must be branched as shown so that they are in contact with the glow plasma on both sides of the cathode, otherwise control of the plasma potential is lost and a high gap resistance results. Molybdenum is used as the cathode material because stable, reproducible electrical characteristics can be obtained with this metal if the cathode is thoroughly cleaned by subjecting it to heavy direct-current glow discharge currents. Also, the maintaining voltage of molybdenum cathodes in helium at the pressure used in these valves is 110 volts, which is several tens of volts lower than the maintaining voltage for a nickel cathode. This lower potential simplifies the circuits used in connexion with the valve. Priming is achieved by the use of radioactive material. Typical characteristics for these valves using helium and neon as gas fillings follow.

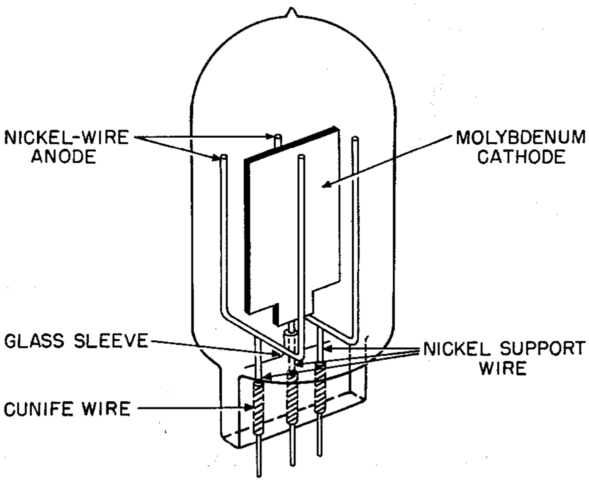

Figure 16—Unit speech-gap valve. The valve envelope is 1·75 inches (4·4 centimetres) tall and 0·354 inch (9 millimetres) in diameter.

Helium-filled valves:—

Direct-current breakdown voltage (independent of glow history) = 240±10 volts

Maintaining voltage = 113± 2 volts

Gap resistance = 100 ohms

Neon-filled valves:—

Direct-current breakdown voltage = 280±20 volts

Maintaining voltage = 105± 2 volts

Gap resistance = 200 ohms

These figures relate to valves where the effect of wall charges upon breakdown voltage is not directly controlled by a connexion to the sputtered metal layer in the inner surface of the envelope. Recent work suggests that such control reduces the spread in V_b and can increase the $V_b - V_m$ gap if the potential at which the sputtered layer is maintained is correctly chosen.

4. *Conclusion*

The operation and characteristics of the speech gap have now been described. Specific embodiments of the ideas presented have been directed mainly towards the utilization of these devices in telephony, but it is clear that there are many other fields in electronic development where their use would be advantageous. It is hoped that this article may stimulate interest in such development.

5. Acknowledgments

The authors wish to acknowledge their debt to their colleagues who have assisted in the design and construction of experimental valves and to the management of Standard Telecommunication Laboratories, Limited, and Standard Telephones and Cables, Limited, for permission to publish this article.

6. Patent References

A. H. W. Beck, T. M. Jackson, and J. Lytollis, British Patent Application 9307/52.

A. H. W. Beck, T. M. Jackson, and J. Lytollis, British Patent Application 24260/53.

Recent Telecommunication Development

"Telecommunications" and "Electronics"

Dr. A. T. Starr, a member of the technical staff of Standard Telecommunication Laboratories is the author of two books recently issued as part of the Pitman Engineering Degree Series. They are companion texts primarily intended for undergraduate students working for bachelor's degrees in engineering at the University of London. They are, of course, suitable for college instruction generally and for private study.

"Telecommunications" is divided into 10 chapters and 9 appendixes under the following titles.

Chapter 1—Communication Systems
Chapter 2—Elements of Telegraph Systems
Chapter 3—Acoustics, Microphones, and Reproducers
Chapter 4—Elements of Telephone Systems
Chapter 5—Lines and Transmission Networks
Chapter 6—Telecommunication Circuits
Chapter 7—Line Telephony
Chapter 8—Basic Principles of Television
Chapter 9—Electromagnetic Waves, Radiation, and Aerials
Chapter 10—Microwave Technique

Appendix 1—Mathematical Formulae
Appendix 2—Steady-State A.C. Theory
Appendix 3—Fourier Analysis
Appendix 4—Power Level: Decibel; Neper
Appendix 5—Operational Calculus: Laplace Transform Method
Appendix 6—Star-Mesh Transformation
Appendix 7—Distortion and Time Delay
Appendix 8—Vectors: Maxwell's Equations
Appendix 9—Waveguide Theory

"Electronics" is divided into 6 chapters and 7 appendixes as follows:

Chapter 1—Physical Fundamentals
Chapter 2—Valves
Chapter 3—Rectification
Chapter 4—Circuit Theory
Chapter 5—Amplifiers, Oscillators, and Detectors
Chapter 6—Electronic Applications

Appendix 1—Mathematical Formulae
Appendix 2—Steady-State A.C. Theory
Appendix 3—Fourier Analysis
Appendix 4—Power Level: Decibel; Neper
Appendix 5—Operational Calculus: Laplace Transform Method
Appendix 6—M.K.S. System: Maxwell's Equations
Appendix 7—Noise

Both books are $5\frac{1}{2}$ by $8\frac{1}{2}$ inches (14 by 22 centimeters) in size and are bound in hard covers.

"Telecommunications" consists of 443 pages of text together with an 8-page index and 9 pages of preliminary material. It sells for 35 shillings in London and $8.75 in New York.

"Electronics" has 388 pages, a 7-page index, and 8 pages of preliminary material. It is priced at 32/6 in London and $7.50 in New York.

The London address of Sir Isaac Pitman & Sons Limited is 39 Parker Street, W. C. 2, and in New York City it is 2 West 45th Street.

Precision Radar Range Calibrator Incorporating Beacon Function*

By R. D. SINISH

Farnsworth Electronics Company, a division of International Telephone and Telegraph Corporation; Fort Wayne, Indian

RANGE is perhaps the most accurate information that can be obtained from present radar systems, and extremely precise test equipment for measuring range is therefore necessary. In addition, the test equipment should be compact, portable, and capable of reliable, repetitive operation by technical personnel. The *AN/UPM-11A* range calibrator to be described differs from previous designs in that it includes internal circuits for calibrating the range of beacon navigational radars as well as ordinary radars.

The calibrator operates essentially as a transponder. Radio-frequency pulses from the radar under test reach the calibrator either by coupling to the directional coupler in the radar system or through an external pickup horn included in the test set. Each input pulse to the calibrator develops radio-frequency pulses that act on the radar system as if they were a series of artificial targets having accurately controlled ranges. In normal radar operation, the return pulses are at the same frequency as the radar transmission; while in beacon operation, the return pulses are at a nominal frequency of 9310 megacycles per second.

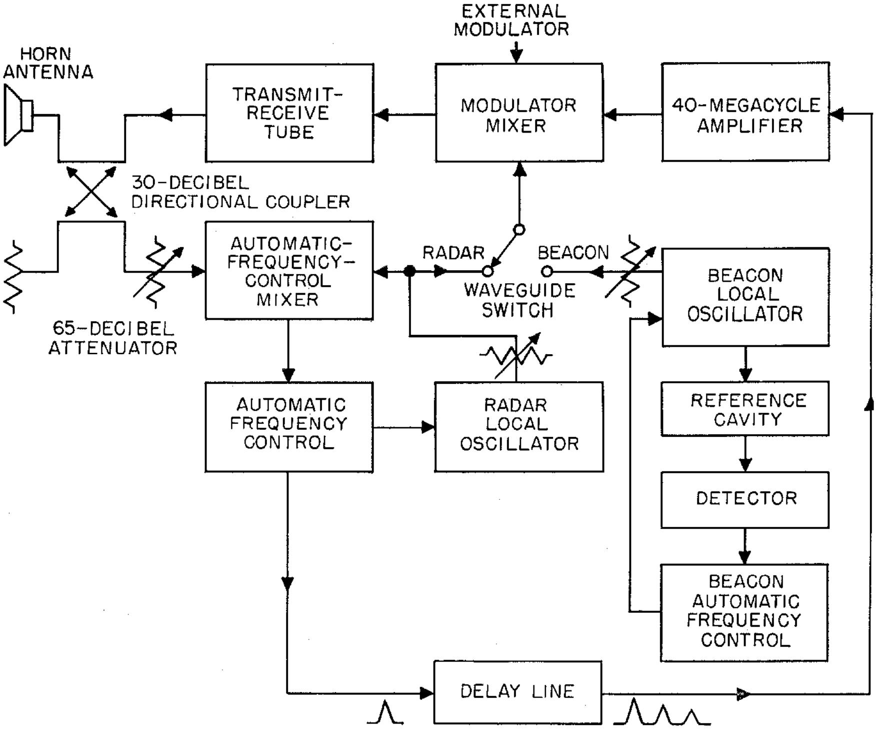

Figure 1—Block diagram of range calibrator.

The calibrator was designed to be operabl with pulsed radar systems in the *X* band, having peak power outputs ranging from 5 to 250 kilowatts, pulse widths of from 0.3 to 3.0 microseconds, and pulse-repetition rates from 300 t 3000 pulses per second. The maximum input radio-frequency power for the required accuracy is 35 microwatts average. At this input level, the return radio-frequency pulses are within 20 kilocycles of the input radio frequency, and the beacon return is within ±2 megacycles of 931 megacycles.

The range accuracy of the first pulse is held to within ±5 yards of the calibrated value and the spacing between subsequent pulses is held to within ±2 yards. This accuracy, both on frequency and pulse spacing, is held over the environmental limits of temperature from −40 degrees to +65 degrees centigrade, relative humidity of 95 percent, vibration of 10*g* from 10 to 55 cycles, and shock of 30*g* for 11 milliseconds.

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* Reprinted in part from "Precision Calibrator Checks Radar Beacons," *Electronics*, volume 28, page 150–153; April, 1955.

The block diagram of the range calibrator is shown in Figure 1. The signal is introduced into the input waveguide through the horn antenna. A portion of this power passes through a 30-decibel directional coupler to the automatic-frequency-control mixer where it is combined with the output of the radar local oscillator.

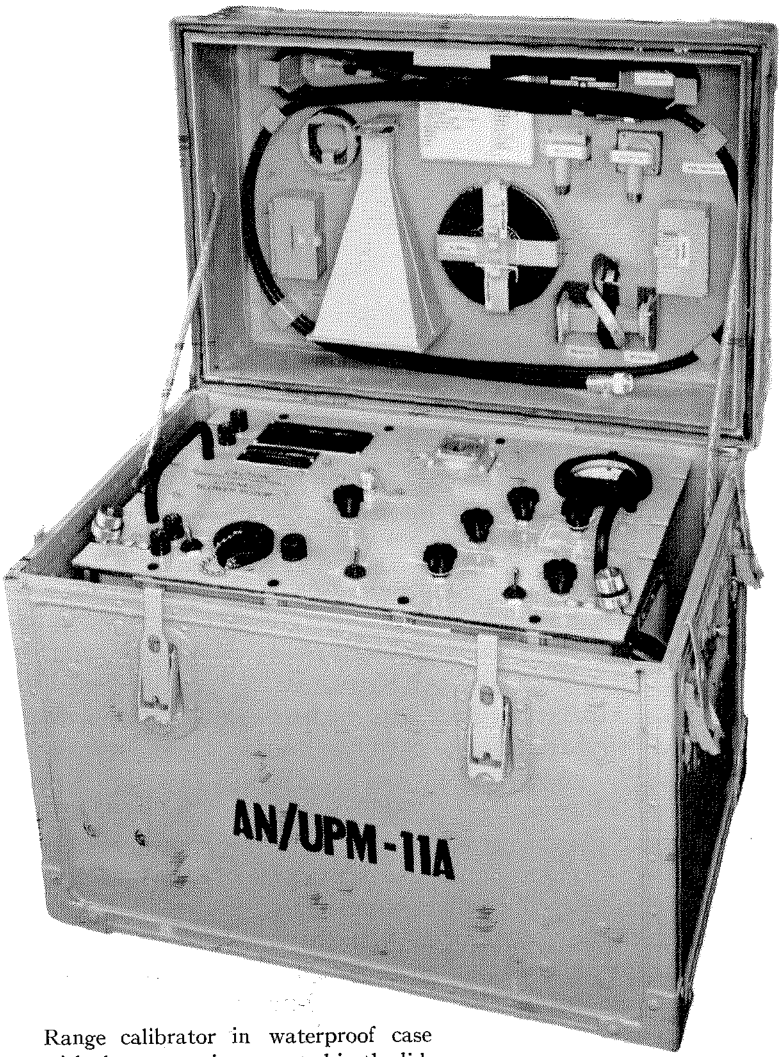
A 65-decibel-maximum attenuator is adjusted to prevent saturation of the crystals in the mixer. Two mixer crystals of type *1N23C* and *1N23CR* connected in reversed polarity permit a push-pull input to the radar automatic-frequency-control chassis.

The automatic frequency control produces an error voltage that maintains the frequency of the radar local oscillator 40 megacycles below the input radio frequency and also furnishes a 40-megacycle pulse to the delay line. The delay line furnishes a series of echo pulses, the first pulse occurring at the approximate one-way delay of the line, with subsequent pulses occurring at intervals of approximately twice the one-way delay of the line. This series of pulses passes through a broad-band amplifier to a *1N23C* crystal in the modulator mixer. The modulator mixer also receives the continuous-wave output of either the radar or beacon local oscillator. The mixing action is such that the pulses go from the mixer through the transmit–receive tube to the radar under calibration. This form of modulation results in two frequencies of output pulses that are 40 megacycles on each side of the local-oscillator frequency. As the local oscillator is held 40 megacycles below the input radar, one pulse output is at the same frequency as the radar while the other pulse output is 80 megacycles below the radar frequency.

On beacon operation, the radar-beacon waveguide switch allows only the beacon local-oscillator output to be injected into the modulator mixer. The beacon local oscillator is held by the beacon automatic frequency control to a nominal frequency of 9350 megacycles, or 40 megacycles higher than the *X*-band beacon frequency. The resultant output pulses from the modulator mixer are thus at 9310 and 9390 megacycles.

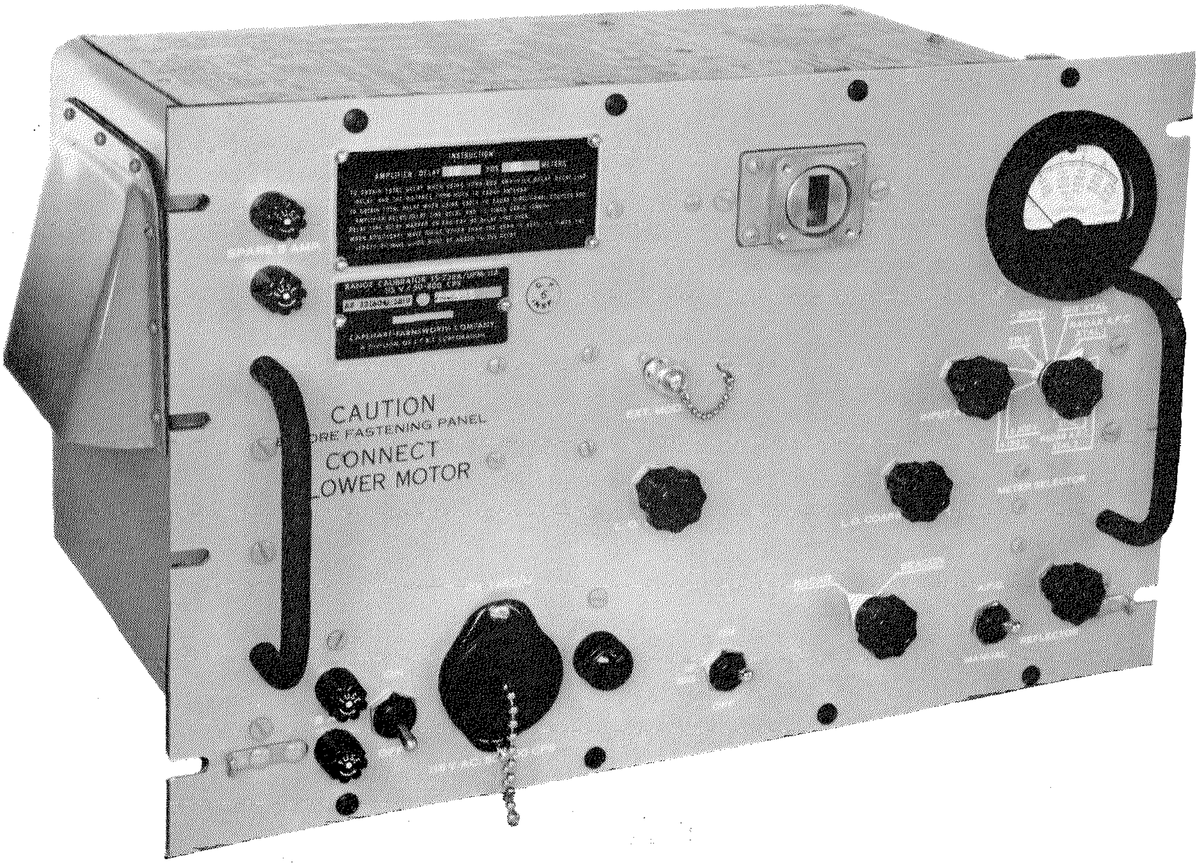
1. Radio-Frequency Assembly

The radio-frequency assembly is fabricated of



Range calibrator in waterproof case with the accessories mounted in the lid.

aluminum. Balanced magic-T mixers are used throughout. A solenoid-operated shutter in the input waveguide assembly provides a 40-decibel insertion loss to protect the mixer crystals when the range calibrator is not in operation. A second attenuator in the input automatic-frequency-control arm, operated by a front-panel control, has a range from 0 to 65 decibels to set the level of the sampled input at a suitable value for the mixer. This control is required with high-power radars to prevent saturation of the crystal mixer and the attendant loss in fidelity of the output pulse from the range calibrator. The radar pulse goes into the E-plane arm of the automatic-frequency-control mixer while the continuous-wave output of the radar local oscillator passes into the H-plane arm of the automatic-frequency-control mixer. Two crystals are mounted in the symmetrical arms of the mixer. The resultant 40-megacycle pulses are carried by a coaxial cable to the radar automatic-frequency-control chassis.



AN/UPM-11A radar range calibrator.

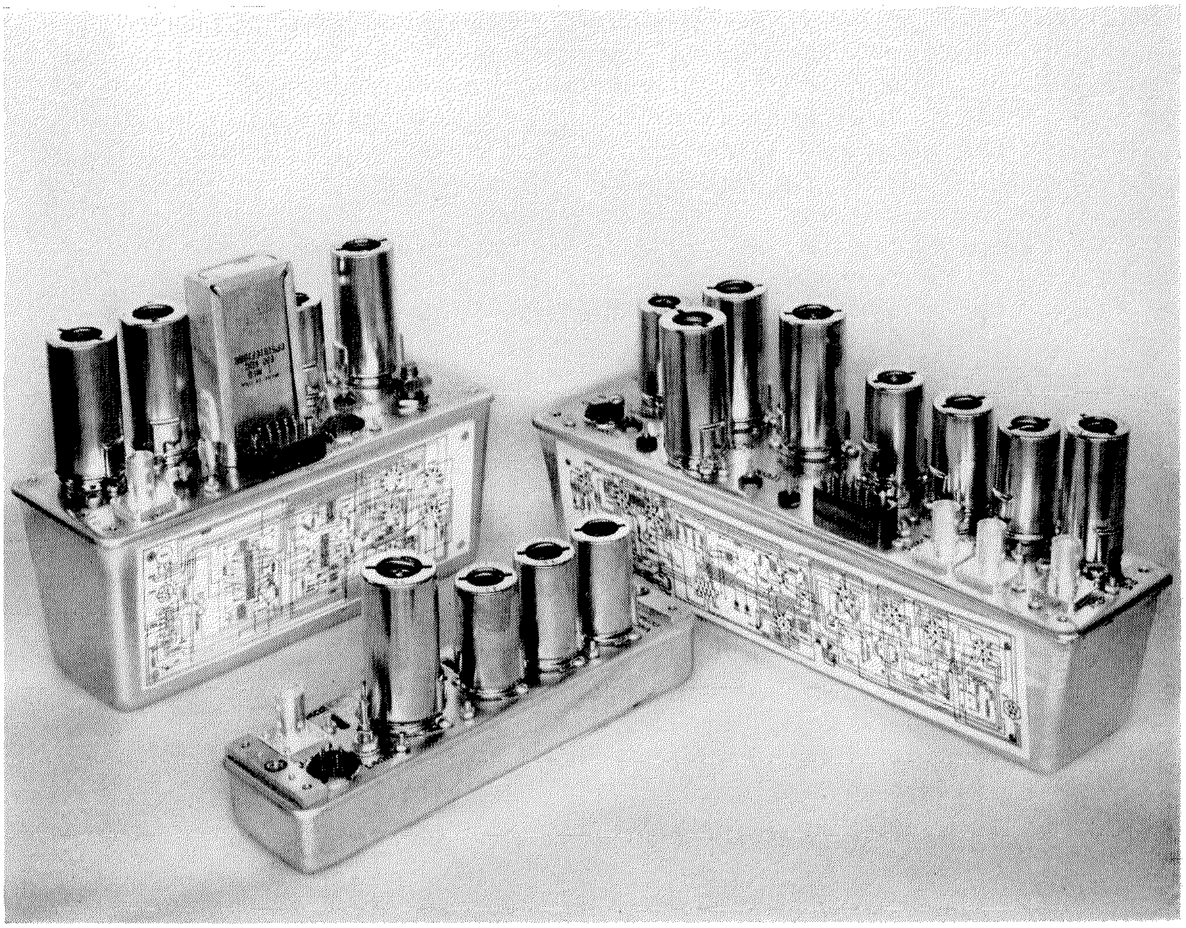
The radar local-oscillator output power goes into the H-plane arm of a modified magic-T mixer. A variable 20-decibel attenuator, adjustable from the front panel, permits setting the output power level. The symmetrical arms of this mixer are connected to the automatic-frequency-control unit and the radar-beacon switch.

The radar-beacon switch contains a movable vane loaded with polyiron. In the radar position, continuous-wave power from the radar local oscillator is supplied to the modulator mixer. In the beacon position, the vane is switched to present a terminating load to the radar local oscillator and at the same time permits continuous-wave power from the beacon local oscillator to go to the modulator mixer. The 30-decibel isolation provided by this shutter prevents interaction between the radar and beacon local oscillators. As a further precautionary step, when the range calibrator is in radar operation, the repeller of the beacon local-oscillator tube is returned to −300 volts, which prevents the tube from oscillating. Removal of the −300 volts by the action of the radar-beacon switch permits instant operation of the beacon local oscillator with no additional warm-up time. This system of isolating the two local-oscillator outputs was adopted to insure that only one output, either radar or beacon, is transmitted to the radar system.

The modulator mixer is a balanced magic-T with the two *1N23C* crystals mounted in the symmetrical arms. Local-oscillator power passes into the *H*-plane arm while the output signals resulting from the mixing action are transmitted through the *E*-plane arm. The 40-megacycle pulses are applied to one of the crystals. The second crystal is connected to a front-panel jack to allow for external modulation. This permits either sine-wave, square-wave, or noise modulation of the output pulses. It is also useful if the calibrator is to produce a simulated one-way target signal in conjunction with an external pulse generator.

A *1B63A* transmit–receive tube is used to protect the modulator-mixer crystals from burnout by high-level input radar pulses.

The output of the beacon oscillator is split in a modified magic-T mixer, one portion going through the radar–beacon switch to the modulator mixer and the other going to the beacon reference cavity. A special beacon reference cavity operating at 9350 megacycles was developed for use in this equipment. The normal frequency tolerance is ±0.3 megacycle while the maximum deviation from nominal over the temperature range is ±0.4 megacycle. A *1N23C* crystal



The 40-megacycle amplifier is flanked on the left by the beacon and on the right by the radar automatic-frequency-control units.

detector is used with the cavity to provide an input error signal to the beacon automatic-frequency-control chassis.

2. Radar Automatic Frequency Control

The radar automatic frequency control is of the basic phantastron sweep type. The two input pulses from the mixer are combined in a broadband degenerate π network and applied to the grids of the cathode-follower and the two-stage intermediate-frequency amplifier shown in Figure 2. The cathode-follower drives the ultrasonic delay line. It has a gain of approximately 0.8 and an over-all band pass of 20 megacycles centered at 40 megacycles. The conventional two-stage intermediate-frequency amplifier operates into a Weiss discriminator. The output *S* curve of the discriminator may be examined at a test point provided for monitoring and for initial alignment. The positive-going peak of the *S* curve occurs at 38.4 megacycles while the negative-going peak occurs at 41.6 megacycles. To prevent overloading of the intermediate-frequency-amplifier stages, automatic-gain-control voltage is derived from the negative-going portion of the discriminator curve. This voltage is amplified in one section of a double triode and applied to the second section of the triode operated at a fixed bias. When the amplified voltage is sufficient to overcome the bias, the excess voltage is filtered and applied to the grids of the two intermediate-frequency-amplifier stages. A test point is provided for monitoring the automatic-gain-control voltage.

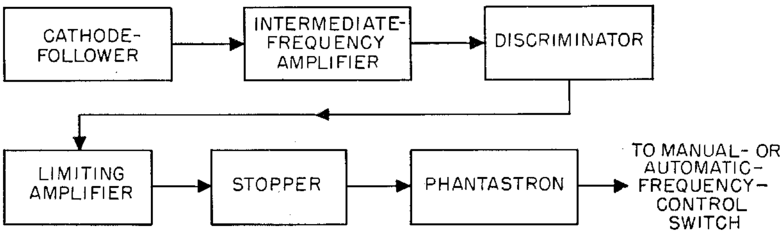

CATHODE-FOLLOWER
INTERMEDIATE-FREQUENCY AMPLIFIER
DISCRIMINATOR
LIMITING AMPLIFIER
STOPPER
PHANTASTRON
TO MANUAL- OR AUTOMATIC-FREQUENCY-CONTROL SWITCH

Figure 2—Radar automatic frequency control.

The output of the discriminator goes to a direct-coupled inverse-freeback-pair video amplifier. This amplifier has high gain at low signal levels and low gain at high signal levels to furnish the required dynamic range. The amplified video signals are connected to a peak-limiting amplifier stage and a stopper tube for the phantastron. The negative-going signal at the output of the limiter tube is connected to the cathode of the stopper tube. The stopper-tube output is filtered and then applied to the input grid of the phantastron.

The phantastron operates as a free-running sawtooth oscillator at a frequency of approximately 2 cycles per second. This sawtooth sweep voltage is applied to the reflector control of the radar local-oscillator tube, where the frequency excursion due to the sawtooth voltage is approximately 60 megacycles. A front-panel control sets the direct-voltage level on the reflector.

The bias voltage from the stopper tube effectively lowers the transconductance of the phantastron to the point where it stops sweeping, at which point the phantastron acts as a direct-current amplifier maintaining the correct voltage level on the radar oscillator. For manual frequency control, −300 volts is applied to the screen of the phantastron to prevent sweeping. At the same time, the voltage on the plate of a thyratron is set to its average level by a voltage-divider network, and the same panel control is used for both automatic and manual operation.

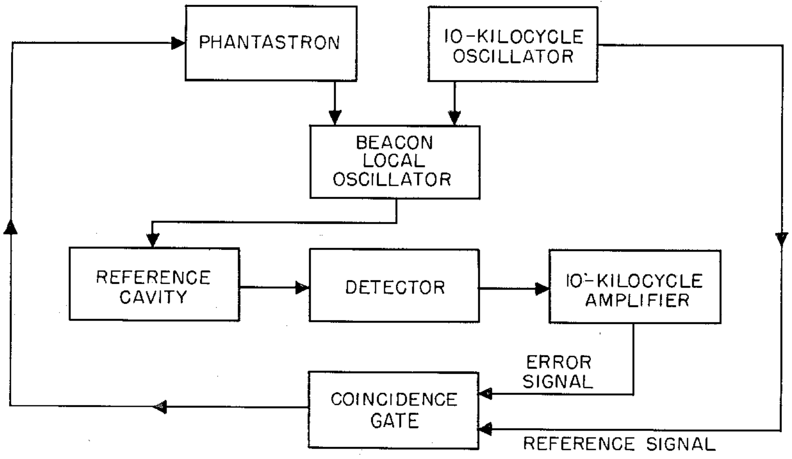
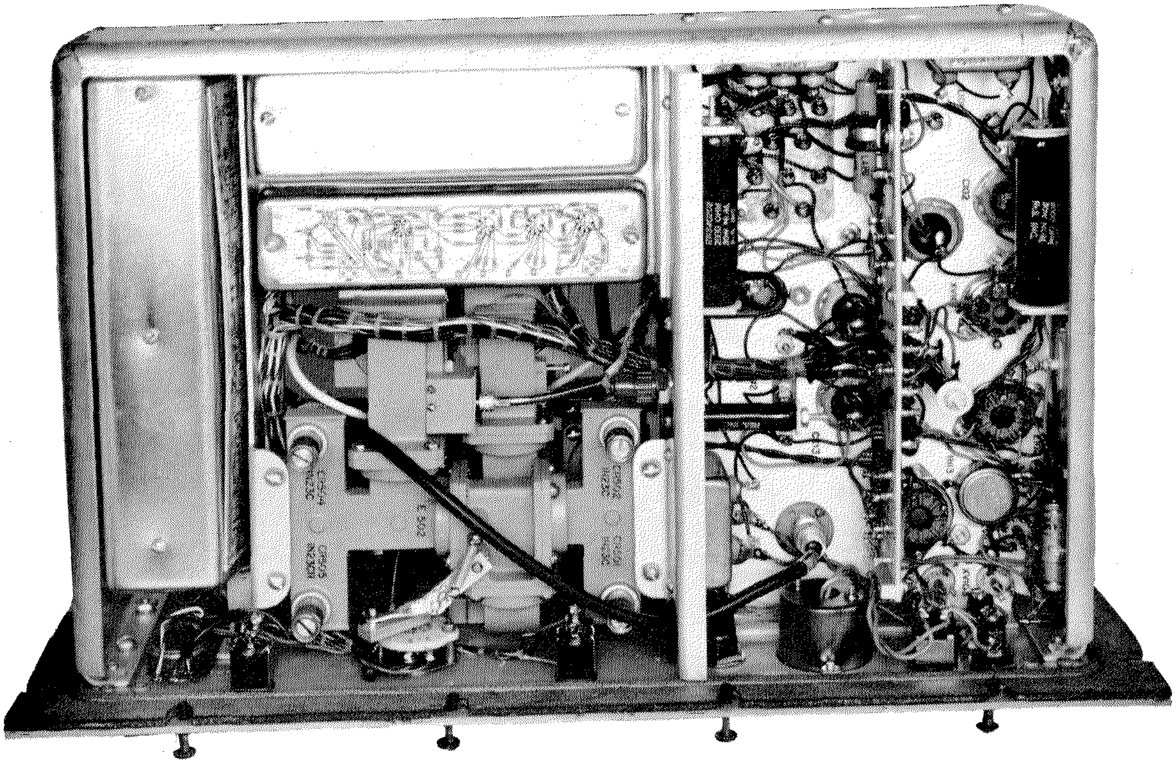


Figure 3—Beacon automatic frequency control.

3. Beacon Automatic Frequency Control

An absolute-frequency type of automatic frequency control is used to maintain the frequency of the beacon local oscillator at 9350 megacycles, the frequency of the beacon reference cavity. Again, as may be seen in Figure 3, a phantastron-type sweep circuit is employed. A 10-kilocycle signal is superimposed on the sweep voltage and slope detection in the cavity is utilized to furnish the frequency-control error voltage.



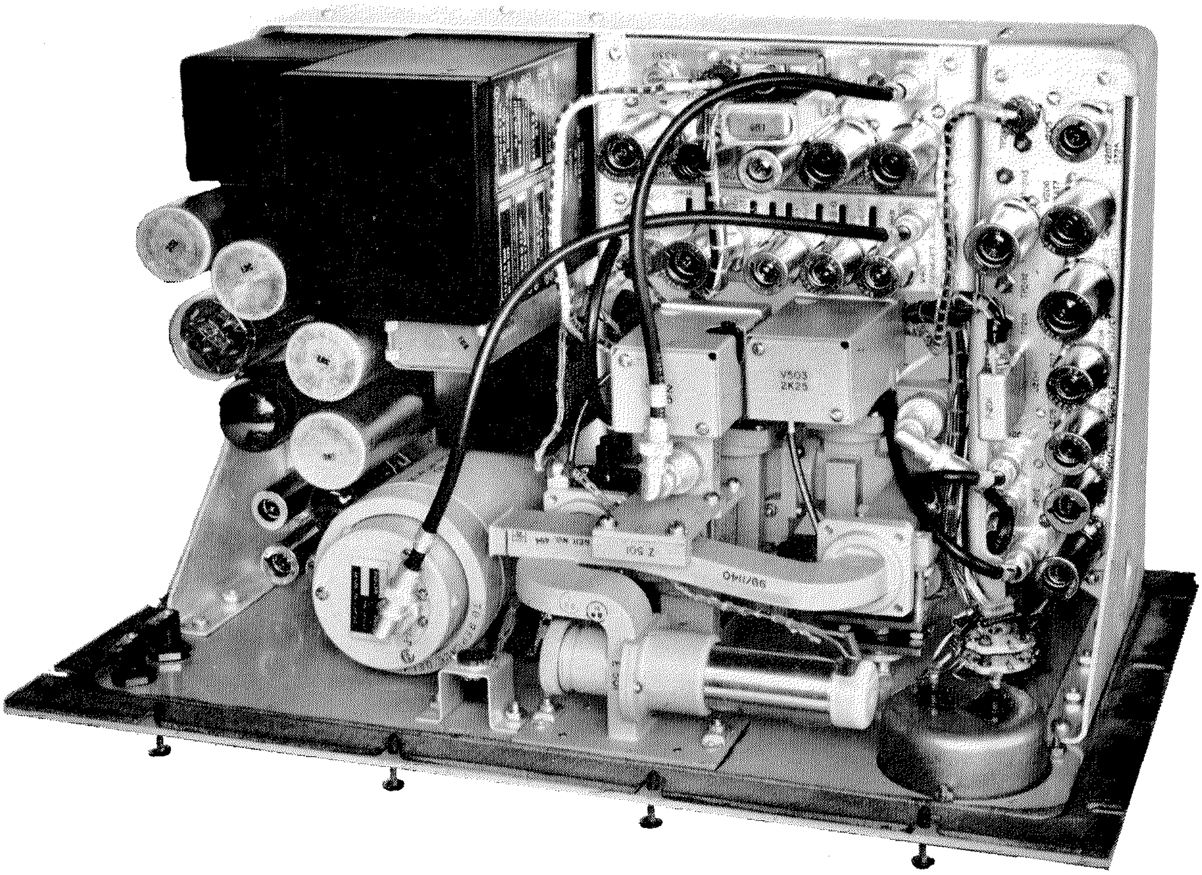
Bottom view of the calibrator with the power supply at the right.

The sawtooth voltage output of the phantastron is modulated with a 10-kilocycle wave and both are applied to the repeller of the beacon local oscillator, thus causing the output frequency of the local oscillator to sweep through the frequency of the beacon reference cavity. The sweep frequency is approximately 13 cycles and the total sweep excursion of the local oscillator is approximately 50 megacycles. As the beacon local oscillator is preset to one frequency, no external controls for this local oscillator are required. Internal adjustments permit setting of the voltage level of the reflector and the injection level.

A 10-kilocycle modulating voltage is taken from a Wien-bridge oscillator and is capacitively coupled to the beacon local-oscillator sweep voltage. A similar reference voltage from that same oscillator goes to one of the control grids of a coincidence tube. A phase-shift network is used to correct for any phase difference between the currents in the plate and cathode resistors from which these two voltages are obtained.

The resulting frequency-modulated output of the beacon local oscillator passes through the beacon reference cavity, which has a minimum Q of 1500. By slope detection in the cavity, the frequency-modulated output is converted to amplitude-modulated power, with the phase of the 10-kilocycle amplitude modulation reversing as the local-oscillator frequency sweeps through the resonant frequency of the cavity. When the beacon local-oscillator frequency is higher than the cavity frequency, the 10-kilocycle modulation is in phase with the cavity output.

The 10-kilocycle modulation is detected and applied to the 10-kilocycle amplifier in the beacon automatic-frequency-control chassis. The over-all gain of this three-stage amplifier is approximately 20 000. The amplified error signal is applied to the other control grid of the coincidence tube. When the error signal and the reference signal are in phase, the coincidence tube conducts, applying a filtered voltage bias that stops the sweep action of the phantastron.



The individual chassis are mounted in a U-shaped bracket that is fastened to the front panel. The radar automatic-frequency-control unit is at the extreme right. Next to it, from the panel outward, are the radio-frequency head, 40-megacycle amplifier, and beacon automatic-frequency-control chassis. The delay-line oven is mounted on the panel in front of the regulated power supply.

4. Delay Line and Oven Assembly

The delay line is a hermetically sealed unilateral unit approximately 1 inch in diameter by 3 inches long (2.5 by 7.6 centimeters). It employs a *Z*- or *X*-cut quartz bar, the physical length of which determines the delay through the line. A 40-megacycle transducer crystal is bonded to each end of the delay bar to convert from 40-megacycle electric waves to ultrasonic waves and back again to electric pulses. The input pulse from the cathode-follower of the radar automatic frequency control is used to shock-excite the delay line. The first pulse is propagated through the line and picked up by the output transducer crystal. A portion of the pulse energy is reflected back through the line to the input transducer crystal, where it is again reflected back to the output crystal; this process is repeated until the energy is dissipated. It is thus apparent that the second and succeeding pulses are spaced at intervals approximately twice the one-way delay of the line. The total number of useful pulses is in the vicinity of 15 to

20 depending on the length of the delay bar. Each succeeding pulse is of smaller amplitude than the preceding pulse, thus effectively adding to the realism of the pulse simulation of remote radar targets. The pass band of the delay line is a minimum of 12 megacycles centered at 40 megacycles. The insertion loss measured at the tenth pulse is approximately 55 decibels down from the input pulse.

To maintain the required over-all stability of the delay line, a thermostatically controlled oven is used. The temperature of this oven is set to 71 ±2 degrees centigrade initially and the maximum temperature deviation is held to ±0.3 degree over the on–off heating cycle. The oven is entirely self-contained with a removable end bell for easy insertion of the delay line. Inductances are included in the oven to tune out the capacitance of the connecting cables and of the delay-line input and output circuits.

5. The 40-Megacycle Amplifier

The 40-megacycle amplifier is necessary to overcome the losses in the ultrasonic delay line. This chassis consists of three pentodes and a twin triode. The circuit used is a staggered triple-tuned design followed by a broadband grounded-grid amplifier preceding the cathode-follower output stage. The amplifier has a 10-megacycle pass band centered at 40 megacycles and an over-all gain of 37 decibels.

It was required to maintain the inherent signal delay in the 40-megacycle amplifier fairly constant over a wide variety of input conditions as this delay is the major portion of the calibrated equipment delay. Design tests showed that the impedance of the signal modulator crystal varied considerably with the injection level of the continuous-wave local-oscillator signal. This variation affected the over-all gain–bandpass characteristics, and therefore the signal delay of the 40-megacycle amplifier. The use of the broadband grounded-grid final amplifier stage with cathode-follower output successfully overcame this difficulty.

6. Mechanical Construction

To increase the serviceability of the range calibrator, the separate circuits are packaged individually. The range calibrator is composed of six individual mechanical assemblies: radar automatic frequency control, beacon automatic frequency control, 40-megacycle amplifier, oven assembly, radio-frequency head assembly, and the power supply. All units are mechanically bound together by a separate U-shaped wraparound band that is tied into the front panel. This permits unit substitution of chassis for rapid maintenance. With the exception of the power supply, all units may be removed by loosening the four screws that tie the units to the assembly. In addition, the shields on the beacon and radar automatic frequency controls and the 40-megacycle amplifier are readily removable for circuit checking with the chassis installed in the calibrator. One main cable is used to distribute power to the various chassis.

The *AN/UPM-11A* consists of a transit case, range calibrator, and accessory equipment that is mounted in the cover of the case. The over-all size of the watertight transit case is 22 inches by 19 inches by 15 inches (56 by 48 by 38 centimeters) and the total weight, including the range calibrator and all accessories is 110 pounds (50 kilograms). The range calibrator is designed for standard relay-rack mounting, having front-panel dimensions of 12¼ inches by 19 inches and extending 13 inches behind the front panel (31 by 48 by 33 centimeters). The range-calibrator unit weighs 60 pounds (27 kilograms). The accessories include the input power cable, 8-foot (2.4-meter) coaxial cable, 2 waveguide-to-coaxial-cable connectors, measuring tape calibrated in yards and meters, waveguide twist section, 2 quick-disconnect waveguide fittings, and a 20-degree pickup horn.

7. Acknowledgments

The development and subsequent production of this range calibrator was done under Air Force Contract AF33(604)-5819. Mr. J. J. Pokorny of the Armament Laboratory, Wright Air Development Center; Dayton, Ohio, served as project engineer for the Air Force. Credit is also due Messrs. E. L. McDirmit, G. A. Richards, D. E. Fisher, R. W. Harpel, R. E. Saxe, and L. D. Stewart, all of Farnsworth Electronics Company, who participated in the development of this equipment.

Cathode-Follower Phase-Shift Oscillator

By J. C. SAMUELS

Farnsworth Electronics Company, a division of International Telephone and Telegraph Corporation; Fort Wayne, Indiana

CATHODE–FOLLOWER amplifiers may be modified by connecting a phase-shifting network between grid and cathode to produce an oscillator. Analysis indicates that this oscillator is inherently more stable than the standard arrangement in which the phase-shifting network is placed in the plate circuit.

The frequency of oscillation and the necessary gain for sustained oscillations are derived for three typical oscillators. Experimental models are discussed briefly. Preliminary measurements indicated frequency to be very insensitive to supply-voltage variations and reasonably independent of alternating-current loading at the cathode.

• • •

Oscillators derived from the cathode-follower amplifier have been studied by Dunn,[1] Bacon and Salmon,[2] and Schlesinger.[3] The oscillators of Dunn and of Bacon and Salmon utilized networks of resistance and capacitance with overunity gain and are therefore properly called resistance–capacitance cathode-follower oscillators. Preliminary work on these circuits indicated that they are very stable with respect to supply-voltage variations. The oscillator described by Schlesinger employed inductance and capacitance for the elements and was termed a Colpitts-type cathode-follower oscillator. No information was given on frequency stability.

The consequences of modifying the cathode-follower amplifier in such a manner as to produce an oscillator that might properly be termed a cathode-follower phase-shift oscillator will be discussed. It is a counterpart of the standard phase-shift oscillator[4] but with the phase-shifting network in the cathode circuit instead of the plate circuit. The placing of the phase-shifting network in the cathode circuit appears to have a few good points that are not realized when the network is in the plate circuit. Recent synthesis[1,2,5,6,7] of networks of only resistance and capacitance with overunity gain suggests their use with resultant advantages in the phase-shifting network of the cathode-follower oscillator.

1. *Analysis*

1.1 Basic Circuits

The standard phase-shift oscillator[4] is shown in Figure 1. Under the assumptions of the linear-amplifier theory, the frequency of oscillation of the phase-shift oscillator is determined from

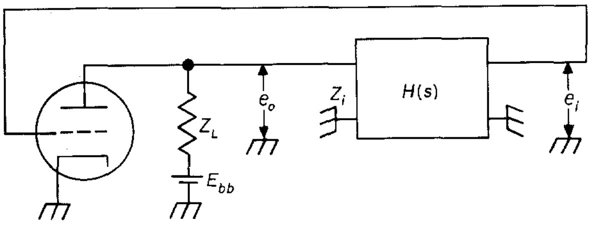


Figure 1—Standard phase-shift oscillator.

$$H(s) + \frac{r_p + Z_0}{\mu Z_0} = 0, \qquad (1)$$

where

$H(s)$ = transfer function of phase-shifting network
s = parameter of the Laplace transformation
r_p = plate resistance
μ = amplification factor
$Z_0 = (Z_L Z_i)/(Z_L + Z_i)$
Z_L = load impedance
Z_i = input impedance of phase-shifting network.

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[1] S. C. Dunn, "R-C Cathode-Follower Feedback Circuits," *Wireless Engineer*, volume 30, page 19; January, 1953.

[2] W. Bacon and D. P. Salmon, "Resistance-Capacitance Networks With Over-Unity Gain," *Wireless Engineer*, volume 30, pages 20–22; January, 1953.

[3] Kurt Schlesinger, "Cathode-Follower Circuits," *Proceedings of the IRE*, volume 33, pages 843–855; December, 1945.

[4] E. L. Ginzton and L. M. Hollingsworth, "Phase-Shift Oscillators," *Proceedings of the IRE*, volume 29, pages 43–49; February, 1941.

[5] C. L. Longmire, "An RC Circuit Giving Over-Unity Gain," *Tele-Tech*, volume 6, pages 40–41; April, 1947.

[6] H. Epstein, "Synthesis of Passive RC Networks with Gains Greater than Unity," *Proceedings of the IRE*, volume 39, pages 833–835; July, 1951.

[7] A. D. Fialkow and I. Gerst, "Maximum Gain of an RC Network," *Proceedings of the IRE*, volume 41, pages 392–395; March, 1953.

The cathode-follower phase-shift oscillator is shown in Figure 2.

The condition for sustained oscillations is

$$1 - H(s)\alpha(s) = 0$$

or for linear operation

$$H(s) - \frac{r_p + (\mu + 1)Z_0}{\mu Z_0} = 0, \qquad (2)$$

where

$H(s)$ = transfer function of phase-shifting network
s = parameter of the Laplace transformation
r_p = plate resistance
μ = amplification factor
$Z_0 = (Z_k Z_i)/(Z_k + Z_i)$
Z_k = cathode load impedance
Z_i = input impedance to phase-shift network
$\alpha(s)$ = cathode-follower gain.

An examination of (1) and (2) shows that for the same tube and comparable Z_0, the cathode-follower phase-shift oscillator should be less dependent on tube characteristics than the standard phase-shift oscillator.

1.2 EXAMPLES OF CATHODE–FOLLOWER OSCILLATORS

Some typical examples of cathode-follower phase-shift oscillators are given with their frequencies of oscillation as determined from linear theory.

1.2.1 Resistance–Inductance–Capacitance

The frequency of oscillation of the circuit shown in Figure 3 is given by

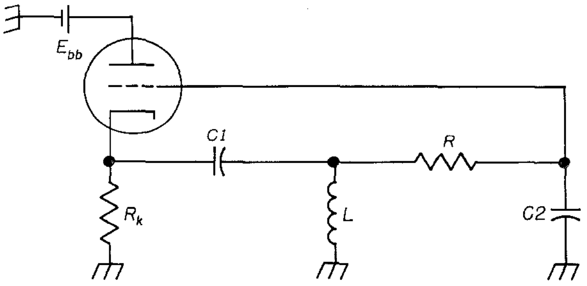

Figure 3—Cathode-follower oscillator using resistance–inductance–capacitance phase-shifting network.

$$f = \frac{1}{2\pi}\left\{\frac{a - 1}{(a - \bar{\alpha})LC_2 + C_2r_1[C_2r_2 - \beta\bar{\alpha}(C_2R_2)] - (C_2R)^2}\right\}^{1/2} \qquad (3)$$

Since tube characteristics are primarily functions of the supply voltages, the cathode-follower phase-shift oscillator should have better frequency stability with respect to supply-voltage variations than the related standard phase-shift oscillator. It would also appear that the frequency of the cathode-follower oscillator would be less dependent on the effects of loading.

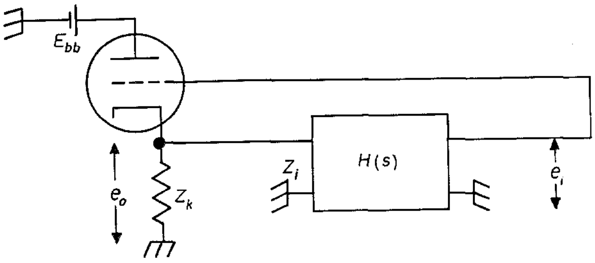

Figure 2—Cathode-follower phase-shift oscillator.

where $\bar{\alpha}$ is a physically realizable solution of

$$\alpha^2 + (\beta_1 + \beta_4 - \beta_3)\alpha + \beta_1\beta_4 - \beta_2\beta_3 = 0. \qquad (4)$$

where

$$\beta_1 = \frac{a(C_2r_1) + C_2r_2 - 2RC_2}{C_2R - C_2r_1 - \beta C_2R_2}$$

$$\beta_2 = -\frac{\mu}{\mu + 1}\frac{r_2}{R_k}$$

$$\beta_3 = \frac{(a-1)(LC_2)(\beta C_2R_k)}{[LC_2 + (\beta C_2R_k)C_2r_1][C_2R - C_2r_1 - \beta C_2R_k]}$$

$$\beta_4 = \frac{aLC_2 + (C_2r_1)(C_2r_2) - (C_2R)^2}{LC_2 + (\beta C_2R_k)(C_2r_1)}$$

$$\beta = \frac{(\mu + 1)}{\mu}$$

$$r_1 = R + R_L$$

$$r_2 = R + R_k$$

$$a = 1 + (C_2/C_1).$$

1.2.2 Colpitts Type

The Colpitts oscillator shown in Figure 4 was considered by Schlesinger. The frequency of oscillation is given by

$$f = \frac{1}{2\pi}\left\{\frac{1}{LC_2} + \frac{1}{LC_1}\left[1 + (1-\alpha)\frac{R_L}{R_0}\right]\right\}^{1/2}, \quad (5)$$

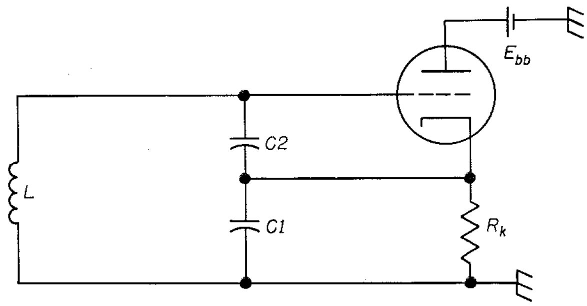

Figure 4—Colpitts-type cathode-follower phase-shift oscillator.

where

$$R_0 = \frac{r_pR_k}{r_p + (\mu+1)R_k}$$

R_L = coil resistance

and α is a physically realizable solution of

$$\frac{\alpha^2}{C_1^2R_0^2} - \left(2\tau_1 + \tau_2 + \frac{\tau_3}{\tau_2}\right)\frac{\alpha}{C_1R_0} + \left(\tau_1 + \frac{\tau_3}{\tau_2}\right)(\tau_1 + \tau_2) - \tau_1 = 0, \quad (6)$$

where

$\tau_1 = 1/(C_1R_0)$
$\tau_2 = R_L/L$
$\tau_3 = 1/(LC_2) + 1/(LC_2)$
α = required cathode-follower gain.

1.2.3 Resistance–Capacitance

In Figure 5 is shown a cathode-follower oscillator using only resistance and capacitance in the phase-shifting network. The corresponding standard phase-shift oscillator is also shown. For equal resistances and equal capacitances, the frequency of oscillation of the cathode-follower circuit of Figure 5 may be calculated from

$$f = \frac{1}{2\pi RC\left[\dfrac{1 + 3R_0/R}{5(1-\alpha) + R_0/R}\right]^{1/2}}, \quad (7)$$

in which

$$R_0 = \frac{r_pR_K}{r_p + (\mu+1)R_k}$$

and α is the required cathode-follower gain given by

$$\alpha = \frac{1}{60}\left\{59 + 23\left(\frac{R_0}{R}\right) \pm \left[1 - 46\left(\frac{R_0}{R}\right) + 49\left(\frac{R_0}{R}\right)^2\right]^{1/2}\right\}. \quad (8)$$

If $R \gg R_0$, then (8) becomes

$$\left.\begin{aligned}\alpha &= 58/60 \text{ (using minus before radical)}\\ &= 59/60 \text{ (using plus before radical)}\end{aligned}\right\} \quad (8A)$$

and using these values in (7) the frequency of oscillation is

$$\left.\begin{aligned}f &= \frac{1}{2\pi RC}\frac{1}{6^{1/2}} \text{ (− sign for radical)}\\ &\text{or}\\ &= \frac{1}{2\pi RC}\frac{1}{2 \times 3^{1/2}} \text{ (+ sign for radical)}\end{aligned}\right\} \quad (7A)$$

It is of interest to compare these results with those for the corresponding standard phase-shift oscillator. It is found[4] that its frequency of oscillation is

$$f = \frac{1}{2\pi RC[6 + 4(R_0/R)]^{1/2}} \quad (9)$$

and the required gain is

$$A = 29 + 23\left(\frac{R_0}{R}\right) + 4\left(\frac{R_0}{R}\right)^2, \quad (10)$$

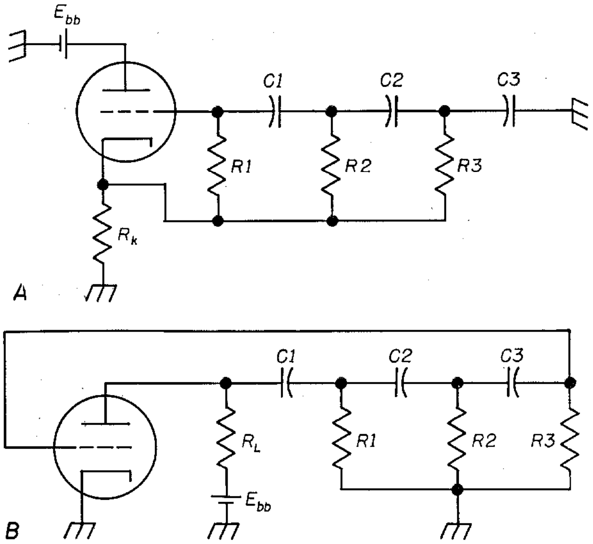

Figure 5—Oscillators using only resistance and capacitance for phase shift. The cathode-follower (grounded-plate) version is at *A* and the standard (grounded-cathode) form is at *B*.

where

$$R_0 = \frac{R_L r_p}{R_L + r_p}.$$

If $R \gg R_0$, then (9) becomes

$$f = \frac{1}{2\pi RC}\frac{1}{6^{1/2}} \tag{9A}$$

and (10) becomes

$$A = 29. \tag{10A}$$

The location of the phase-shifting network in the cathode circuit should result in an oscillator more stable than the standard phase-shift oscillator because plate-resistance variations are reduced by the factor $1/(\mu + 1)$.

2. *Experimental*

In accordance with the principles set forth in the preceding portions of this paper, the following oscillators were set up in breadboard form and investigated briefly.

2.1 EXPERIMENTAL RESISTANCE–INDUCTANCE–CAPACITANCE

The circuit of Figure 6 had a frequency of oscillation of 4550 cycles per second for $L = 200$ millihenries. It produced a good sine-wave output having a root-mean-square value of about 0.25 volt with 200 volts of plate supply. There was a slight change (less than 0.5 percent) of frequency with a supply-voltage variation from 180 to 270 volts. The long-time frequency stability was good. There was about 0.033-percent change of frequency for a 10-percent change of filament voltage, and about 0.062-percent change for a similar change in plate voltage. A tenfold increase (150 000 to 15 000 ohms) of cathode load produced 1.5-percent change of oscillation frequency.

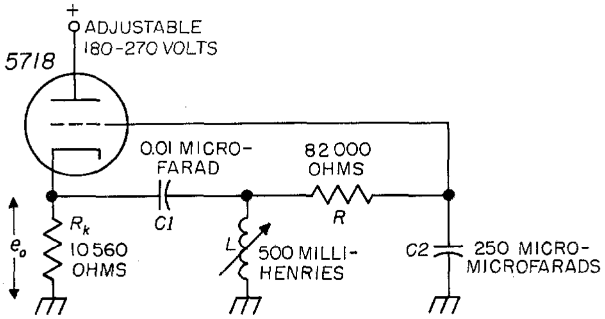

Figure 6—Experimental cathode-follower oscillator using a phase-shifting network consisting of resistance, inductance, and capacitance. The heater was operated with alternating current. The plate supply had negligible alternating-current impedance.

By varying R, L, $C1$, and $C2$, the oscillator was made to cover the frequency range from 41 to 6500 cycles.

2.2 EXPERIMENTAL COLPITTS TYPE

The Colpitts oscillator of Figure 7, which is a modification of the cathode-follower amplifier given in reference 3, was found to give a strong distorted output under the conditions indicated.

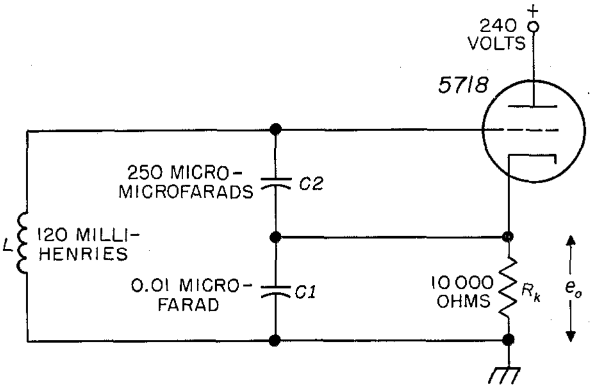

Figure 7—Experimental Colpitts oscillator. The heater was operated with alternating current.

The frequency of oscillation was 24 500 cycles, and the root-mean-square voltage output across the cathode resistor was 6.5 volts. No attempt was made to adjust circuit parameters for best output waveform. The waveform can be considerably improved by adjusting the gain of the cathode-follower so as just to maintain oscillations.

2.3 EXPERIMENTAL RESISTANCE–CAPACITANCE

Figure 8 shows a cathode-follower oscillator using only resistances and capacitances for phase shifting.

Grading of the elements in the phase-shifting network was necessary to reduce the required cathode-follower gain. The use of equal elements in a phase-shifting network of three sections demands a gain of about 58/60 out of the cathode-follower amplifier. This gain is very difficult to obtain with existing vacuum tubes.

The oscillator of Figure 8 had a frequency of oscillation of 137 cycles with a root-mean-square output at the cathode of 20.5 volts. The waveform of the output was approximately sinusoidal.

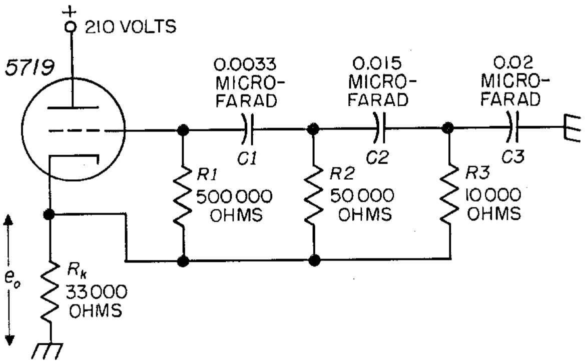

Figure 8—Experimental cathode-follower oscillator using a phase-shifting network of resistances and capacitances The heater was operated with alternating current.

No accurate measurements were made of frequency stability with respect to plate-voltage variations, but there are indications that less than 0.1 percent of frequency change results for a 10-percent increase in plate voltage at some plate voltages. From no load to a cathode load of 43 000 ohms, the oscillation frequency changed from 137 to 161 cycles while the root-mean-square output voltage dropped from 20.5 volts to 7.5 volts.

3. *Concluding Remarks*

The cathode-follower oscillator of the phase-shift type is very stable with respect to supply-voltage variations and is reasonably stable with changes in cathode loading. The location of the phase-shifting network in the cathode circuit should result in an oscillator inherently more stable than the standard phase-shift oscillator. Further development work is required to establish this point more definitely.

Telephone Statistics of the World*

FOR THE SIXTH consecutive year, 5 million telephones were added throughout the world, making a total of more than 89 million on January 1, 1954. More than half of of the world's telephones were in the United States, where some 5100 privately owned and operated systems gave service to almost one out of every three individuals. In Europe, which had some 28 percent of the world's telephones, mostly under public operation, there was a telephone for roughly one out of every 25 individuals.

For the purpose of this compilation, only those telephones that can be connected to a commercial public system are counted. Eleven countries reported more than 1 million telephones in service on January 1, 1953; United States, United Kingdom, Canada, Western Germany, France, Japan, Sweden, Italy, Australia, Switzerland, and Argentina. Of of the world's principal countries, 8 had more than 15 telephones per 100 of the population: United States (31.3), Sweden (27.7), Canada (24.0), Switzerland (21.9), New Zealand (21.9), Denmark (18.7), Australia (16.1), and Norway (15.8).

New York, with more telephones than any other city, had twice as many as Greater London, which ranked second. On a per-capita basis, Washington, District of Columbia, led among the world's large cities with 62.8 telephones per 100 population and Stockholm, Sweden, was first outside the United States with 51.9.

A subdivision in certain of the tables shows the number of telephones operated under private and government ownership. The latter category has reference to municipal and state, as well as national, ownership. Although the American Telephone and Telegraph Company and its subsidiaries operated about 80 percent of the United States' 50 372 972 telephones on January 1, 1954, there were over 5000 other privately owned companies that furnished service in the United States.

The statistics in this compilation are based on questionnaires sent to the telephone administrations of the various countries throughout the world.

Telephones in Continental Areas

Partly estimated; statistics reported as of other dates have been adjusted to January 1, 1954

Continental Area	Total Telephones			Privately Owned		Automatic (Dial)	
	Number	Percent of Total World	Per 100 Population	Number	Percent of Total	Number	Percent of Total
North America	54 000 600	60.5	30.6	53 460 400	99.0	41 603 500	77.0
Middle America	670,000	0.8	1.2	597 700	89.2	502 000	74.9
South America	2 245 500	2.5	1.9	1 054 500	47.0	1 808 200	80.5
Europe	25 400 700	28.5	4.3	3 982 700	15.7	18 703 800	73.6
Africa	1 181 200	1.3	0.5	22 200	1.9	819 300	69.4
Asia	3 662 200	4.1	0.3	2 729 800	74.5	1 696 700	46.3
Oceania	2 039 800	2.3	14.6	143 500	7.0	1 348 100	66.1
World	89 200 000	100.0	3.6	61 990 800	69.5	66 481 600	74.5
United States	*50 372 972*	*56.5*	*31.3*	*50 372 972*	*100.0*	*39 100 000*	*77.6*

* Abridgement from a booklet issued by the American Telephone and Telegraph Company; New York, New York.

TELEPHONES IN COUNTRIES OF THE WORLD AS OF JANUARY 1, 1954

Country	Total Telephone	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
NORTH AMERICA					
Alaska	23 533	14.71	74.3	6 608	16 925
Canada	3 603 900	24.01	69	3 080 780	523 120
Greenland	0	—	—	—	—
St. Pierre and Miquelon	208	4.52	0	0	208
United States	50 372 972	31.27	77.6	50 372 972	0
MIDDLE AMERICA					
Bahamas	5 866	6.82	99.3	0	5 866
Barbados	5 148	2.31	97.3	5 148	0
Bermuda	7 600	18.54	100	7 600	0
British Honduras	854	1.17	0	45	809
Canal Zone (1) (2)	6 965	24.02	100	0	6 965
Costa Rica	10 600	1.18	0	10 300	300
Cuba	141 055	2.38	89.8	141 055	0
Dominican Republic	8 359	0.36	81.6	8 218	141
El Salvador	15 000	0.72	75	0	15 000
Guadeloupe	1 362	0.50	0	0	1 362
Guatemala	6 428	0.21	84.7	0	6 428
Haiti	4 182	0.12	91	0	4 182
Honduras	7 000	0.44	60	0	7 000
Jamaica	18 020	1.20	97.4	18 020	0
Leeward Islands	902	0.75	0	0	902
Martinique	3 047	1.10	66.2	0	3 047
Mexico	330 221	1.16	72.1	328 675	1 546
Netherlands Antilles	6 905	3.79	96.7	0	6 905
Nicaragua (3)	3 500	0.30	0	0	3 500
Panama	16 182	1.85	76.9	15 711	471
Puerto Rico	47 367	2.11	52.7	43 909	3 458
Trinidad and Tobago	18 997	2.80	86.3	18 997	0
Virgin Islands (United States)	1 700	6.30	0	0	1 700
Windward Islands					
Dominica	307	0.54	0	0	307
Grenada	875	1.07	0	0	875
Saint Lucia	395	0.47	0	0	395
Saint Vincent	422	0.60	0	0	422
Total	1 999	0.68	0	0	1 999
SOUTH AMERICA					
Argentina	1 001 158	5.39	79.8	75 430	925 728
Bolivia	11 110	0.35	94.3	11 110	0
Brazil	679 540	1.22	82.1	679 540	0
British Guiana	3 908	0.87	7.5	0	3 908
Chile	145 139	2.36	69	144 649	490
Colombia	128 970	1.05	87.6	0	128 970
Ecuador	11 500	0.33	60.9	1 500	10 000
Falkland Islands	319	13.87	0	0	319
French Guiana	336	1.16	0	0	336
Paraguay (3)	5 800	0.38	86.2	0	5 800
Peru	58 017	0.63	83.4	58 017	0
Surinam	2 741	1.19	91.1	0	2 741
Uruguay	104 510	4.05	75	0	104 510
Venezuela	92 420	1.70	93.7	84 240	8 180
EUROPE					
Albania (3)	1 700	0.14	0	0	1 700
Andorra (3)	100	2.00	0	0	100
Austria	458 006	6.58	80	0	458 006
Belgium	777 340	8.84	77.2	0	777 340
Bulgaria (3)	61 000	0.82	43	0	61 000
Channel Islands					
Guernsey and Dependencies	9 568	20.80	0	0	9 568
Jersey	12 174	21.36	0	0	12 174
Total	21 742	21.11	0	0	21 742
Czechoslovakia (4)	350 708	2.88	59.4	0	350 708
Denmark	825 879	18.73	39.6	726 144	99 735
Finland	408 531	9.81	62.2	350 700	57 831
France	2 768 951	6.45	64.6	0	2 768 951
Germany, Democratic Republic (3)	250 000	1.42	60	0	250 000
Germany, Federal Republic (5)	3 255 971	6.61	89.8	0	3 255 971
Gibraltar	1 399	6.08	89.5	0	1 399
Greece	104 237	1.32	93.7	0	104 237
Hungary (3)	122 000	1.27	74	0	122 000

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(1) Excluding telephone systems of the military forces.
(2) June 30, 1953.
(3) Data partly estimated.
(4) January 1, 1948 (latest official statistics).
(5) March 31, 1954.
(6) January 1, 1936 (latest official statistics). On January 1, 1954, nine cities connected.
(7) Includes the Isle of Man, but not the Channel Islands.

TELEPHONES IN COUNTRIES OF THE WORLD AS OF JANUARY 1, 1954—*Continued*

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
EUROPE—*Continued*					
Iceland	23 774	15.85	64.4	0	23 774
Ireland	103 798	3.52	68.1	0	103 798
Italy	1 774 462	3.74	93.7	1 774 462	0
Liechtenstein	2 334	16.67	100	0	2 334
Luxemburg	28 150	9.26	72.5	0	28 150
Malta and Gozo (5)	7 869	2.46	0	0	7 869
Monaco	6 020	28.67	100	0	6 020
Netherlands	919 572	8.70	91.6	0	919 572
Norway (2)	530 827	15.79	61.5	70 400	460 427
Poland (3)	240 000	0.93	67	0	240 000
Portugal	208 143	2.40	57.4	133 842	74 301
Rumania (3)	141 000	0.85	60	0	141 000
Saar	44 938	4.66	99.6	0	44 938
San Marino (3)	300	2.31	0	0	300
Spain	903 097	3.15	80.3	888 334	14 763
Sweden	1 994 378	27.73	71	838	1 993 540
Switzerland	1 074 216	21.90	98.1	0	1 074 216
Trieste	36 648	9.99	97	35 898	750
Union Soviet Socialist Republics (6)	861 181	0.52	19.9	0	861 181
United Kingdom (5) (7)	6 139 229	12.15	75.3	0	6 139 229
Yugoslavia (3)	149 000	0.87	70	0	149 000
AFRICA					
Algeria	116 889	1.23	75.7	0	116 889
Anglo-Egyptian Sudan	11 648	0.13	76.5	0	11 648
Basutoland	509	0.09	4.9	0	509
Bechuanaland Protectorate	131	0.04	0	0	131
Belgian Congo and Ruanda-Urundi	11 451	0.07	62.6	0	11 451
British East Africa					
Kenya	20 407	0.35	72.3	0	20 407
Tanganyika	7 949	0.10	55.3	0	7 949
Uganda	6 328	0.12	77.6	0	6 328
Total	34 684	0.18	69.4	0	34 684
British West Africa					
Gambia	386	0.14	99.2	0	386
Gold Coast	9 581	0.21	37.1	0	9 581
Nigeria	15 063	0.05	38.5	0	15 063
Sierra Leone	1 573	0.08	84	0	1 573
Total	26 603	0.07	41.6	0	26 603
Comoro Islands	0	—	—	—	—
Egypt	135 388	0.62	79.5	0	135 388
Ethiopia	4 776	0.03	74.6	0	4 776
French Cameroon	2 258	0.07	0	0	2 258
French Equatorial Africal	1 851	0.04	43.5	0	1 851
French Somaliland	570	0.90	0	0	570
French Togo	611	0.06	0	0	611
French West Africa	17 860	0.10	39.7	0	17 860
Liberia	487	0.02	100	0	487
Libya	6 055	0.49	80.6	0	6 055
Madagascar	7 650	0.17	46.8	0	7 650
Mauritius	5 881	1.12	6	0	5 881
Morocco					
French Zone	80 651	1.01	77.6	0	80 651
Spanish Zone	8 076	0.78	56.4	8 076	0
Tangier Zone	9 546	5.19	97.4	9 297	249
Total	98 273	1.07	77.8	17 373	80 900
Nyasaland Protectorate	2 407	0.10	86.2	0	2 407
Portuguese Africa					
Angola	2 268	0.05	85.5	0	2 268
Cape Verde Islands	126	0.08	0	0	126
Mozambique	6 236	0.11	76.6	0	6 236
Portuguese Guinea	257	0.05	91.1	0	257
South Tome and Principe	277	0.46	0	0	277
Total	9 164	0.08	75.9	0	9 164
Rhodesia, Northern	6 077	0.30	94.5	0	6 077
Rhodesia, Southern	37 063	1.62	78.7	0	37 063
Réunion	3 557	1.35	0	0	3 557
Saint Helena	89	1.78	0	0	89
Seychelles and Dependencies	50	0.14	0	0	50
Somalia	856	0.06	0	0	856
Somaliland Protectorate	225	0.03	0	0	225
South West Africa	6 733	1.57	49.2	0	6 733
Spanish Guinea	525	0.26	67.4	525	0
Spanish North Africa	4 320	3.02	58.1	4 320	0

TELEPHONES IN COUNTRIES OF THE WORLD AS OF JANUARY 1, 1954—*Continued*

Country		Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
					Private	Government
AFRICA—*Continued*						
Spanish West Africa						
Ifni		114	0.32	0	0	114
Spanish Sahara		52	0.10	0	0	52
Total		166	0.19	0	0	166
Swaziland		566	0.28	1.9	0	566
Tunisia		30 666	0.84	59.4	0	30 666
Union of South Africa	(5)	606 152	4.55	68.4	0	606 152
Zanzibar and Pemba	(3)	1 100	0.40	0	0	1 100
ASIA						
Aden Colony		1 550	1.55	100	0	1 550
Aden Protectorate		0	—	—	—	—
Afghanistan		5 822	0.05	30	0	5 822
Bahrain		1 045	0.93	100	1 045	0
British Borneo						
Brunei		125	0.25	0	0	125
North Borneo		971	0.28	66.9	0	971
Sarawak		720	0.12	0	0	720
Total		1 816	0.18	35.8	0	1 816
Burma		7 136	0.04	0	0	7 136
Cambodia		1 884	0.05	0	0	1 884
Ceylon		22 855	0.28	94	0	22 855
China (not including Formosa)	(4)	244 028	0.05	72.9	94 945	149 083
Cyprus		8 145	1.60	79.4	8 145	0
Formosa		34 586	0.41	47	0	34 586
French India		78	0.02	100	0	78
Hong Kong	(2)	40 434	1.84	100	40 434	0
India	(5)	210 868	0.06	50.7	1 966	208 902
Indonesia		63 977	0.08	3.8	0	63 977
Iran		39 300	0.19	57.3	0	39 300
Iraq	(5)	28 010	0.50	67.5	0	28 010
Israel		47 430	2.84	89.3	0	47 430
Japan	(5)	2 594 506	2.96	40.2	2 594 506	0
Jordan		6 536	0.47	75.6	0	6 536
Korea, South		28 461	0.13	31.7	0	28 461
Kuwait		988	0.66	84.8	988	0
Laos		286	0.02	0	0	286
Lebanon		24 368	1.81	79.7	0	24 368
Malaya		40 259	0.69	57.1	1 250	39 009
Maldive Islands		0	—	—	—	—
Muscat and Oman		137	0.02	100	137	0
Netherlands New Guinea		772	0.11	0	0	772
Pakistan		27 886	0.04	56	0	27 886
Philippine Republic		41 807	0.19	64.4	34 753	7 054
Portuguese Asia						
Macao		1 789	0.89	99.5	0	1 789
Portuguese India		224	0.03	0	0	224
Portuguese Timor		312	0.07	0	0	312
Total		2 325	0.18	76.6	0	2 325
Qatar		105	0.58	100	105	0
Ryukyu Islands						
Okinawa		1 500	0.23	0	0	1 500
Others		300	0.10	0	0	300
Total		1 800	0.19	0	0	1 800
Saudi Arabia		7 743	0.13	0	0	7 743
Singapore		28 895	2.52	100	28 895	0
Syria		27 155	0.75	86.5	0	27 155
Thailand		8 020	0.04	100	0	8 020
Turkey		113 609	0.50	79.4	0	113 609
Viet-Nam		13 980	0.06	42.8	0	13 980
Yemen		0	—	—	—	—
Other Places	(3)	32 000	0.16	25	0	32 000
OCEANIA						
American Samoa		323	1.79	100	0	323
Australia		1 432 776	16.07	65.5	0	1 432 776
French Oceania						
French Settlements		710	1.13	0	0	710
New Caledonia and Dependencies		1 830	2.82	0	0	1 830
Total		2 540	1.98	0	0	2 540
Guam		4 412	5.81	90.7	0	4 412
Hawaii		143 461	27.48	96.1	143 461	0
Nauru		0	—	—	—	—
New Hebrides Condominium		108	0.22	0	0	108
New Zealand	(5)	456 289	21.86	59.6	0	456 289

TELEPHONES IN COUNTRIES OF THE WORLD AS OF JANUARY 1, 1954—*Continued*

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
OCEANIA—*Continued*					
Pacific Islands (British)					
Fiji Islands	3 210	1.00	0	0	3 210
Gilbert and Ellice Islands	113	0.30	74.3	84	29
Pitcairn Island	0	—	—	—	—
Solomon Islands Protectorate	0	—	—	—	—
Tonga (Friendly) Islands	245	0.46	0	0	245
Total	3 568	0.69	2.4	84	3 484
Pacific Islands (United States Administration)					
Caroline Islands	177	0.42	0	0	177
Mariana Islands (less Guam)	0	—	—	—	—
Marshall Islands	457	4.15	0	0	457
Total	634	1.07	0	0	634
Papua and New Guinea	2 943	0.20	0	0	2 943
Western Samoa	551	0.61	0	0	551

TELEPHONE CONVERSATIONS FOR THE YEAR 1953

Conversation data were not available for all countries

Country	Number of Conversations in Thousands			Conversations Per Capita
	Local	Toll	Total	
Algeria	59 000	20 400	79 400	8.5
Argentina	3 106 400	39 100	3 145 500	171.1
Australia	997 700	78 400	1 076 100	121.9
Belgium	444 000	69 700	513 700	58.5
Brazil	2 182 900	38 200	2 221 100	39.8
Canada	5 948 000	131 900	6 079 900	411.3
Ceylon	54 800	3 700	58 500	7.2
Chile	383 500	22 400	405 900	66.8
Colombia	444 200	4 800	449 000	37.3
Cuba	811 200	5 200	816 400	140.0
Denmark	978,200	166 000	1 144 200	261.7
Egypt	271 900	10 900	282 800	12.9
Finland	509 200	77 300	586 500	141.7
France	1 444 900	467 800	1 912 700	44.7
Germany, Federal Republic (1)	2 077 900	447 900	2 525 800	51.6
Greece	271 900	4 600	276 500	35.2
Iceland	54 900	1 400	56 300	375.3
Ireland	80 300	12 700	93 000	31.6
Israel	74 100	3 500	77 600	47.2
Italy	2 787 000	177 200*	2 964 200	62.7
Jamaica	44 000	700	44 700	30.3
Japan (1)	8 761 000	568 600	9 329 600	107.4
Luxemburg	10 300	8 600*	18 900	62.6
Malaya	117 900	11 800	129 700	22.7
Mexico	749 300	9 800	759 100	27.1
Morocco	71 400	15 300	86 700	8.7
Norway (2)	465 200	53 200	518 400	154.9
Peru	211 600	2 900	214 500	23.7
Philippine Republic	280 000	400	280 400	13.3
Portugal	189 800	34 700	224 500	26.0
Puerto Rico	94 300	3 200	97 500	43.5
Saar	55 600	1 800	57 400	59.5
Spain	2 249 000	73 300	2 322 300	81.4
Sweden (3)	2 096 700	123 900	2 220 600	308.8
Switzerland	437 400	358 400*	795 800	162.9
Tunisia	16 600	7 300	23 900	6.7
Turkey	159 500	5 500	165 000	7.4
Union of South Africa (1)	731 100	47 000	778 100	59.0
United Kingdom (1) (4)	3 404 200	279 600	3 683 800	72.9
United States	59 355 000	2 185 000	61 540 000	385.3
Uruguay	306 600	4 600	311 200	123.2
Venezuela	388 600	13 600	402 200	73.9

(1) Year ended March 31, 1954.
(2) Year ended June 30, 1953.
(3) Year ended June 30, 1954.
(4) Includes the Isle of Man, but not the Channel Islands.
* Three-minute units.

United States Patents Issued to International Telephone and Telegraph System February - April, 1955

UNITED STATES patents numbering 28 were issued between February 1 and April 30, 1955 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that are of more-than-usual interest are included.

R. P. Arthur and E. L. Earle, Electrical Terminal Pin Block, 2 703 394.

R. M. Barnard and S. B. Buckley, Metallizing the Surface of Ceramic Bodies, 2 706 682.

A. Brown and T. A. Benner, Panel Control Key, 2 702 321.

M. den Hertog, Two-Stage Group-Selector Circuit, 2 706 748.

L. A. de Rosa, Radio Detection System, 2 703 401.

S. H. M. Dodington, Pulse Repeaters, 2 706 773.

H. Grayson, T. S. McLeod, and V. Vernon-Smith, Frequency-Modulation Carrier-Current System, 2 703 865.

P. W. Hemminger and J. I. Bellamy, Switching System using Capacitor Storage of Digits, 2 700 071.

K. Klinkhammer, Group Selector of Relay Systems, 2 705 743.

H. R. Kough, Electrical Terminal and Terminal Assembly, 2 707 274.

J. S. Macmullan and A. Mortlock, Variable Limiter and Noise Suppressor for *RK9* System, 2 706 776.

F. P. Mason and A. T. Watts, Facsimile Telegraph System, 2 705 739.

W. Master, Dry-Contact Rectifier, 2 707 251.

B. McAdams, Adaptor Circuit, 2 707 211.

K. Muhmer, Contact Spring Assembly, 2 705 602.

A. J. Mullarkey, Axle Counting Arrangements, 2 701 301.

H. L. Overman, Circuit-Component Selecting Unit, 2 706 252.

H. Pletscher, Twinplex Telegraph Signal Receiver, 2 701 276.

W. H. P. Pouliart and F. M. Michiels, Receiving Circuit Arrangement, 2 704 361.

S. J. Powers and W. F. Bonner, Germanium Diode, 2 703 856.

N. Prietzel, Central Switch for Pneumatic Dispatch Systems, 2 703 687.

N. H. Saunders, Conference Provision for Telephone Switchboards, 2 700 072.

K. Seiler, Method of Making Surface-Type and Point-Type Rectifiers, 2 701 216.

K. Seiler and L. Fedotowski, Semiconductor Rectifier or Amplifier of any Desired Surface Profile, 2 702 361.

A. T. Starr and G. King, Switches for High-Frequency Waves, 2 705 776.

V. J. Terry, Electronic-Telegraph Relay Circuits, 2 701 277.

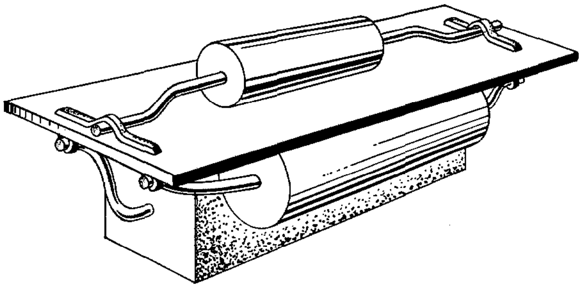
B. V. Thompson and E. W. Deiss, Electrical Coupling Apparatus, 2 707 271.

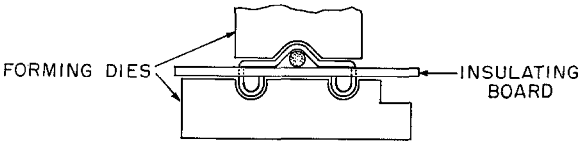
C. deB. White and K. A. Matthews, Amplifier Employing Semiconductor, 2 701 281.

Electrical Terminal and Terminal Assembly

H. R. Kough
2 707 274—April 26, 1955

An electrical terminal in the form of a staple-like member with a bridge portion having a single connecting loop on one side of an insulating sheet, the ends of the staple being curled toward the sheet to provide two connection loops of approximately semicircular outline.





Adaptor Circuit

B. McAdams
2 707 211—April 26, 1955

An adaptor circuit for interconnecting voice-frequency telephone lines with a carrier-frequency system. This development automatically converts the direct-current pulses of the normal dialing and supervisory signals into carrier supervisory signals by means of simple switch connections applying different potentials from separate sources to dial trunk leads under control of dial signals.

Receiving Circuit Arrangement

W. H. P. Pouliart and F. M. Michiels
2 704 361—March 15, 1955

A receiving circuit for use in magnetic-drum storage systems, in which the waveform output alternates to give positive and negative variations. Separate circuits respectively accept the positive and negative parts of the output, each circuit having a short-time-contant storage circuit. Connections between the charging path of each storage circuit and the discharging path of the other storage circuit are included to reduce spurious signals that may occur.

Method of Making Surface-Type and Point-Type Rectifiers and Crystal-Amplifier Layers from Elements

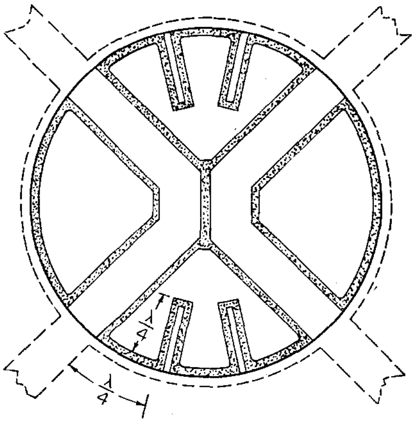
K. Seiler
2 701 216—February 1, 1955

A method of forming semiconductors for point-contact rectifiers and amplifiers to avoid excess impurities that may cause short-circuits in the region adjacent to the point contacts but to preserve adequate conductivity in adjacent conducting zones. The method includes making a vapor deposition of the semiconductor on a base-plate, the conductor substance having a mixture of acceptor or donor impurities and reducing the proportion of these impurities during deposition.

Switches for High-Frequency Waves

A. T. Starr and G. King
2 705 776—April 5, 1955

A switch for waveguides having a rotor carrying passages to connect to different fixed waveguides and having an arrangement to eliminate the need of spring contacts between the rotor and fixed waveguides. The switch includes slots cut radially into the rotor drum from the outer circumference, the slots being positioned equal distances from the openings of the passageways and a peripheral distance from these openings of approximately a quarter wavelength.

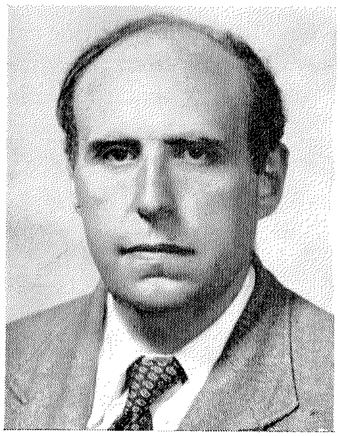


Contact-Spring Assembly for Automatic Switching of Message Carriers in Pneumatic-Tube Systems

K. Muhmer
2 705 602—April 5, 1955

A pneumatic-tube system in which the message carriers are provided with marking rings set to convey a particular message as to destination or route to be followed, with a detecting arrangement for reading the message consisting of contact springs insulated from one another and held out of contact with the rings. A common actuating device for the contact includes a leaf spring and an actuating device for the leaf spring depresses the insulated springs into contact with the rings when the carrier is properly positioned for contact.

Contributors to This Issue



A. H. Beck

A. H. Beck was born in Norfolk, England, in 1916. He received a B.Sc. degree in engineering from University College, London.

In 1937, he joined the research staff of Henry Hughes and Sons. From 1941 to the end of the war, he was seconded to the Admiralty Signals Establishment and part of his theoretical work there resulted in the publication of his book on "Velocity Modulated Thermionic Tubes."

In 1947, he left Hughes to establish the Enfield valve laboratory of Standard Telephones and Cables, and is now head of the valve division of Standard Telecommunication Laboratories. His interests include microwave valves, special-purpose gas tubes, and vacuum physics applied to valves. His most-recent book is on "Thermionic Valves."

Mr. Beck is an Associate Member of the Institution of Electrical Engineers and a Member of the Institute of Radio Engineers.

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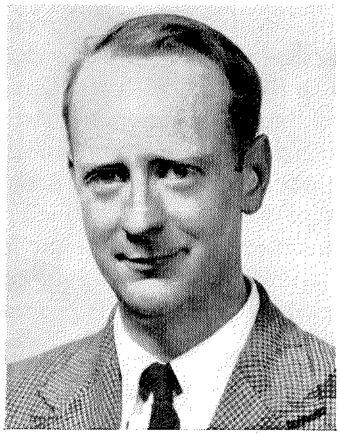

A. D. Brisbane

A. D. Brisbane was born in 1920 and received his early education at Bournemouth. In 1955, he obtained the City and Guilds Final Technological Certificate at Enfield College.

In 1938, he joined the Royal Air Force and after technical training was commissioned in 1942.

After demobilization, he entered the valve laboratory of Standard Telephones and Cables at Enfield, being transferred later to Standard Telecommunication Laboratories. His activities have chiefly concerned thermionic emission.

Mr. Brisbane received the Distinguished Flying Medal in 1941.

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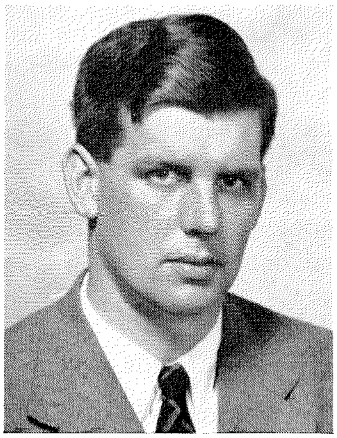


A. B. Cutting

A. B. Cutting was born in Stamford, England, in 1919. He obtained B.Sc. degrees in general science and mathematics from Birkbeck College of the University of London.

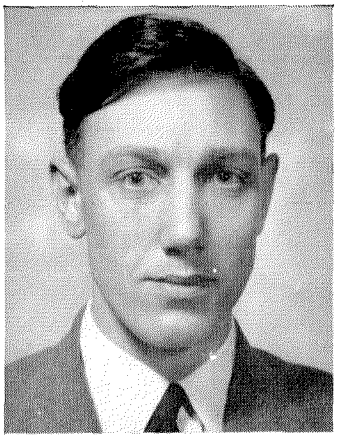
After nine years in the research laboratories of the General Electric Company, he joined the research staff of the Enfield valve laboratory of Standard Telephones and Cables in 1947. He was transferred to Standard Telecommunication Laboratories in 1954. His research work has been primarily on microwave valves.

• • •

J. L. Goodwin was born at Chigwell, England, in 1920. He studied electrical engineering at the Borough Polytechnic and Twickenham Technical College, receiving the Higher National Certificate in 1947.

During the war, he did research work on underwater acoustics with the Admiralty and in 1948 he joined the staff

J. L. Goodwin



T. M. JACKSON

of the transmission division of Standard Telephones and Cables. Mr. Goodwin reports in this issue on a loudspeaker intercommunicating system.

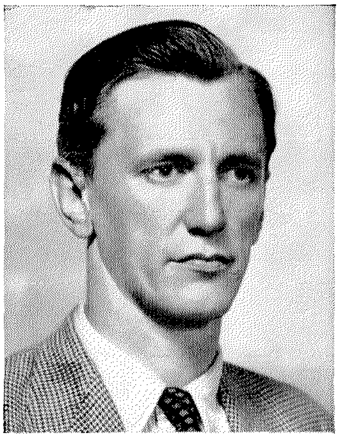
Mr. Goodwin is an Associate Member of the Institution of Electrical Engineers.

• • •

T. M. JACKSON was born in Merioneth, Wales, in 1921.

He joined the Royal Air Force in 1940 and spent four years in experimental work on radar systems at the Government Telecommunications Research Establishment.

In 1946, he came to the Enfield valve laboratories of Standard Telephones and Cables and was transferred to Standard Telecommunication Laboratories in 1954. He has been mainly concerned with devices based on electrical discharges in gases.



G. KING

• • •

G. KING was born in 1925 in London. After technical training at the Royal Technical College in Glasgow, he served the Royal Air Force as a radar fitter.

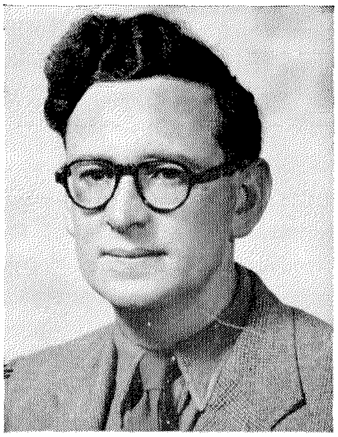
Later he obtained the Final Technological Certificate in telecommunication engineering from the City and Guilds of London.

On demobilization in 1947, he went to the Enfield valve laboratory of Standard Telephones and Cables and in 1954 was transferred to the engineering staff of Standard Telecommunication Laboratories. He is coauthor of the article in this issue on cathodes.

• • •

J. LYTOLLIS was born in 1924 in Skegness, Lincolnshire. He was educated at Worcester College, Oxford, and received a B.A. in physics from the Honour School of Natural Science in 1945.

After two years with the Philips group of companies and three years as an Instructor Officer in the Royal Navy, in 1951 he was employed by Standard Telephones and Cables in the Enfield valve laboratory. He was transferred to the valve division of Standard Telecommunication Laboratories in 1951. His work has been principally on gaseous discharge devices.



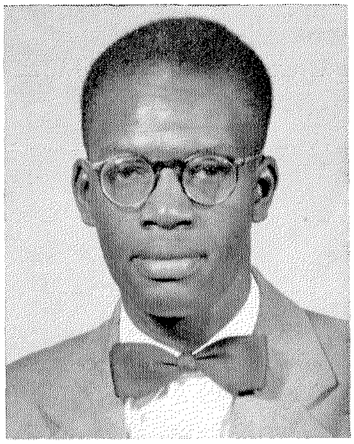
J. LYTOLLIS

• • •

J. C. SAMUELS received the B.S. degree in electrical engineering from Polytechnic Institute of Brooklyn in 1948 and the M.S. degree in mathematics and physics from New York University in 1950. During the next two years, he did graduate work in mathematics at Case Institute of Technology. While at Case Institute and for two years thereafter, he worked on jet-propulsion systems for the National Advisory Committee for Aeronautics.

In 1952, he joined the engineering staff of Capehart–Farnsworth Company and is now with Farnsworth Electronics Company. He reports here on cathode-follower oscillators.

Mr. Samuels is a registered professional engineer in Indiana. He is a



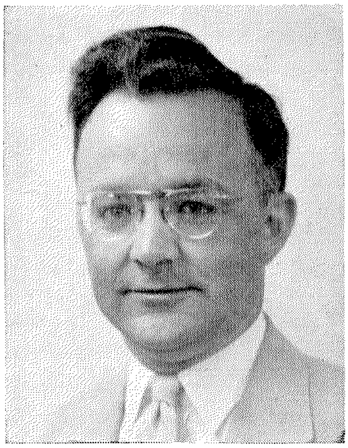
J. C. SAMUELS

member of the Institute of Radio Engineers and a member of Tau Beta Pi and Eta Kappa Nu, both honorary societies.

• • •

R. Donald Sinish was born in Leavenworth, Kansas, on May 2, 1919. He received a B.S. degree in electrical engineering from Purdue University in 1942, where he also did graduate work in high-voltage generation and transmission.

From 1942 to 1944, he was employed by the General Electric Company on the design of high-voltage and special-purpose test equipment. In 1944, he joined the Farnsworth Television and Radio Corporation as field engineer in charge of installation and calibration of precision bombing equipment. His design experience includes radio and radar transmitters and receivers as well as precision test equipment for radar systems.



R. Donald Sinish

Mr. Sinish is a Senior Member of the Institute of Radio Engineers, a Member of the American Institute of Electrical Engineers, and a registered professional engineer.

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Technical Journal of the
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and Associate Companies

MICROWAVE RELAY SYSTEM BETWEEN SAINT JOHN AND HALIFAX

RADIO LINK BETWEEN PUERTO RICO AND THE VIRGIN ISLANDS

MICROWAVE TELEVISION RADIO RELAY SYSTEM

BIHELICAL TRAVELING-WAVE TUBE WITH 50-DECIBEL GAIN

SAMPLING PROCEDURES ON FINISHED CHASSIS AND EQUIPMENTS

UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM

Volume 32 DECEMBER, 1955 Number 4

ELECTRICAL COMMUNICATION

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International Telephone and Telegraph Corporation
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Volume 32 DECEMBER, 1955 Number 4

CONTENTS

Microwave Relay System Between Saint John and Halifax 215
By H. C. Sheffield

Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands 238
By Roger McSweeny

Microwave Television Radio Relay System 248
By O. H. Appelt, Karl Christ, and Kurt Schmid

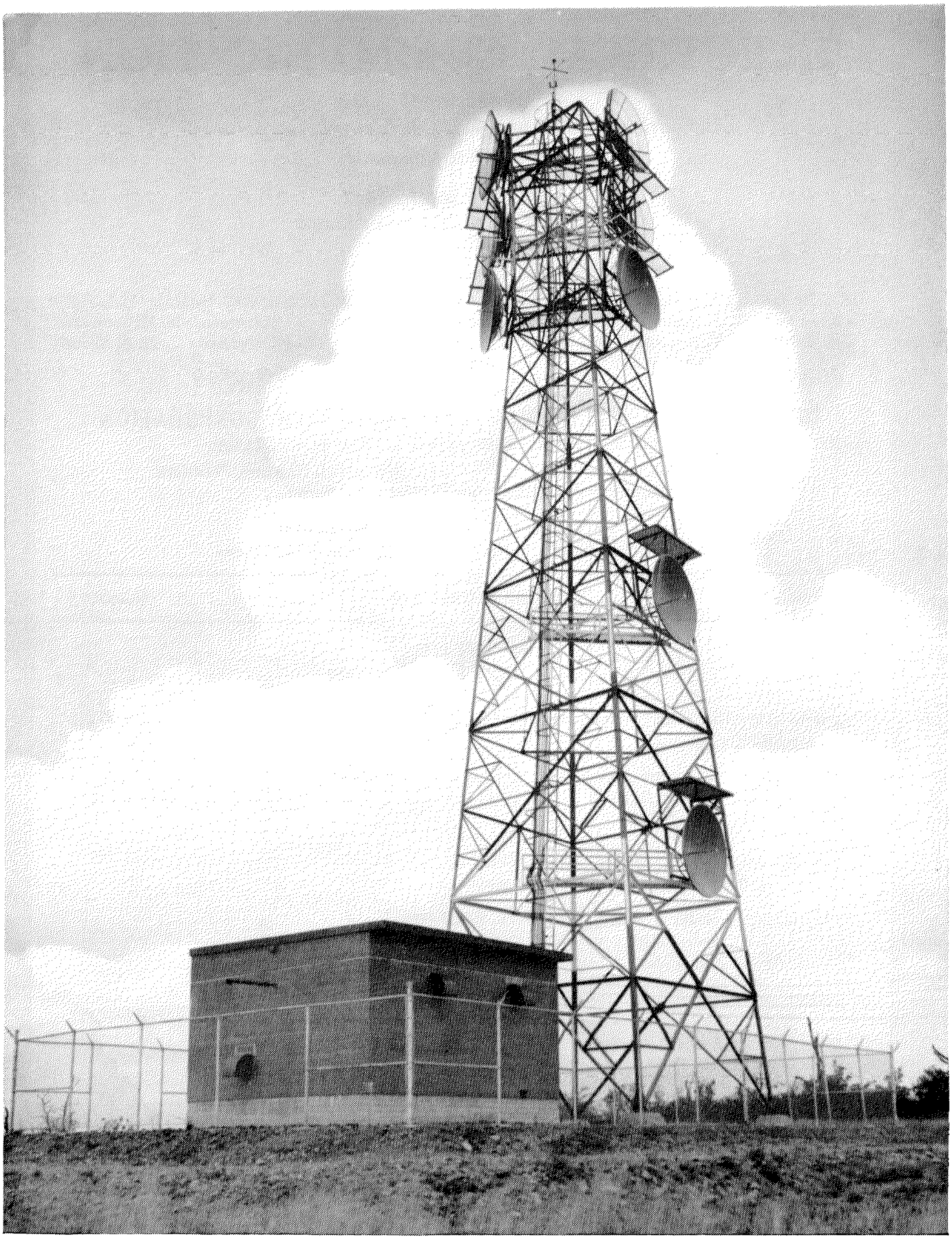
Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles 255
By W. P. G. Klein

Sampling Procedures on Finished Chassis and Equipments 263
By Henry Knapp and Charles Hartmann

United States Patents Issued to International Telephone and Telegraph System; May–July, 1955 271

Recent Telecommunication Developments—
- "Transistors and Other Crystal Valves" 237
- "Electronique Industrielle" 270

Contributors to This Issue 274



Repeater building and self-supporting tower at Otter Lake. The transmission lines are mounted on the safety hoops around the ladder. The screens above the paraboloids are for protection from falling ice.

Microwave Relay System Between Saint John and Halifax*

By H. C. SHEFFIELD

Standard Telephones & Cables Manufacturing Company (Canada) Limited; Montreal, Canada

A MICROWAVE relay system employing time-division multiplex has been constructed to provide toll telephone circuits for the New Brunswick Telephone Company and the Maritime Telegraph and Telephone Company. This system links Saint John in New Brunswick and Halifax in Nova Scotia via five intermediate unattended repeaters and yields 46 four-wire circuits of toll quality that provide out-of-province connections to Montreal, Toronto, Boston, and New York. This system has been in partial use since December 1952, and the complete system has been in continuous service since December 1954. A description of the radio relay link is given primarily from the system-planning aspects. Performance figures are outlined.

1. *Preliminary Planning*

In 1948, the Maritime Telegraph and Telephone Company, which provides the bulk of the long-distance and local circuits in the province of Nova Scotia, was faced with the problem of providing a considerable increase in out-of-province telephone circuits. These circuits would form part of the trans-Canada telephone system, which is operated by the Trans-Canada Telephone Association comprised of seven major Canadian telephone companies (the Maritime and New Brunswick companies are two of these) to provide coast-to-coast telephone service in Canada.

The New Brunswick Telephone Company was planning in conjunction with other telephone companies the installation of two *J*-type carrier systems from Saint John to Montreal and another from Saint John to Bangor, Maine. Many of these circuits were to be extended to Halifax to provide much-needed direct circuits between Halifax and Montreal, Toronto, Boston, and New York. These requirements, coupled with a large increase in traffic between Halifax and Saint John, made imperative the provision of additional circuits between those two cities. Since Halifax had been established as the toll switching centre for Nova Scotia in the national toll-dialing scheme, all long-distance circuits are routed through this point.

The existing circuits were provided by *C*-type carriers operating on open wire pole lines. The pole line followed roughly the path through Moncton and Truro indicated on the map of Figure 1. It is necessarily long to circumvent the Bay of Fundy and was subject to severe icing conditions in winter particularly where it crossed the Isthmus of Chignecto. Severe damage had been experienced several times.

Consideration was therefore given to a microwave relay system running from Halifax across the Annapolis valley and then spanning the Bay of Fundy to Saint John. Such a system would be much shorter than the route around the bay thus allowing a decrease in toll fares. In addition, all maintenance would be confined to a few repeater points and the entire system would be essentially invulnerable to weather conditions. Preliminary investigations indicated that the cost of a microwave system should be appreciably less than that of rebuilding and transposing the pole-line leads to allow the operation of 12-channel carrier systems. Accordingly, engineers of the Maritime Telegraph and Telephone Company, Federal Telecommunication Laboratories, and Federal Electric Manufacturing Company, Limited (now Standard Telephones and Cables Manufacturing Company (Canada) Limited) conducted field surveys of possible repeater points in Nova Scotia in the autumn of 1948.

2. *Propagation Surveys*

Since accurate contour maps of this area were available and weather conditions were generally such as to encourage stable propagation, propagation surveys were considered necessary only over the Bay of Fundy.

This particular 50-mile (80-kilometre) path is nearly all over water. The Bay of Fundy is famous for its prodigious tides, which are as high as 40 feet (12.2 metres). This constant churning action brings the cold water to the surface, frequently causing dense fogs extending several hundred feet above the water. Without doubt, severe temperature inversions also occur. Fortunately, elevations of about 800 feet (245 metres) are available on each side of the bay so that the use of vertical space diversity is practicable. Furthermore, the minimum clearance for these elevations is slightly more than two Fresnel zones, thus allowing considerable abnormal refraction before the path becomes non-optical.

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A propagation survey was made in 1950. Previous experience had shown that the most-abnormal propagation could be expected in the spring, and therefore the tests were conducted during April, May, and June. Bloomsbury Mountain in New Brunswick was chosen as the

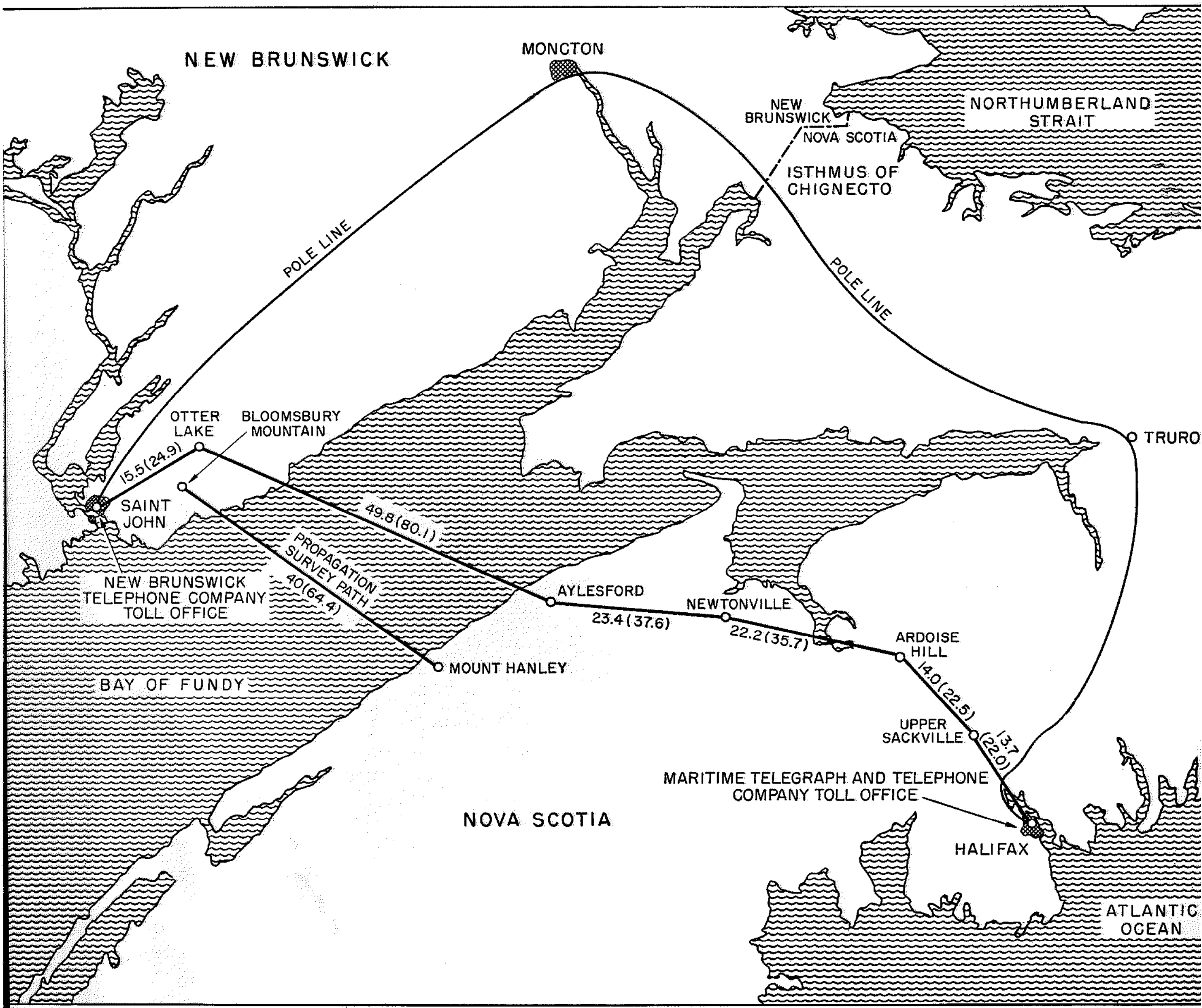

Figure 1—Map showing original pole-line path and radio relay system between Saint John and Halifax. Distances are in miles (kilometres).

likely site for the proposed repeater on the north coast of the Bay of Fundy and was used as the transmitting location. A 6-foot (1.83-metre) paraboloidal antenna, later replaced by a 10-foot (3.1-metre) unit, was installed on a 35-foot (10.7-metre) pole.

In Nova Scotia, the receivers were installed on Mount Hanley. Although it was realized that this would not be the repeater site in the permanent microwave system, it was selected mainly for convenience since it was the terminal of an existing very-high-frequency link. A special 90-foot (27.5-metre) guyed tower was constructed using spliced poles arranged in the form of an A. One receiving dish was located at the top of this and a second one 52 feet (15.9 metres) lower. The signal strengths received by each receiver were continuously recorded as well as their sum. The frequency of the signal was 2000 megacycles per second.

The lower antenna was moved up and down and it was found that the optimum diversity spacing under normal propagation conditions was 52 feet (15.9 metres) as calculated.

The results of the survey indicated that diversity reception was perfectly practicable over this path. Although the fades on individual receivers were frequent, rapid, and very deep (at least 40 decibels), the best signal did not fall more than 10 decibels below the mean for 95 percent of the time. For the remaining 5 percent of the time, fades as deep as 25 decibels were experienced. Based on these results and the use of a microwave time-division system, it was predicted that the signal-to-noise ratio would not be worse than 46 decibels (a considered outage value) for 0.02 per cent of the time, nor worse than 58 decibels for more than 0.2 per cent of the time. All these ratios are based on the measurement of noise in a derived audio channel using an *F1A*-line-weighting curve.

It was observed that the deeper fades occurred on days when poor propagation might be expected, and it was logical to assume that they were caused by abnormal refraction that could upset the diversity spacing. This assumption was further borne out by the fact that the fog layer, which sometimes was prevalent over the Bay of Fundy, was not sufficiently high to intercept the direct ray. Under these conditions, the reflected ray conceivably could suffer severe refraction, while the direct ray suffered only normal refraction. To alleviate this condition, a third antenna could be placed at a distance of one-half the normal diversity spacing below the lower antenna. This would guard against an increase or decrease in the correct diversity spacing for complementary operation.

A propagation survey was also conducted for two days between Bloomsbury Mountain and the roof of the telephone building in Saint John, the proposed terminal site. This path was grazing but could not be improved since the additional height that would be required in Saint John could not be provided by a tower that could be erected on the telephone building. Although it was decided to utilize this grazing path, it was later found necessary for other reasons to move the Bloomsbury Mountain repeater to Otter Lake, thus obtaining a path with full Fresnel-zone clearance.

3. *Repeater Site and Path Planning*

It was decided in February 1951 to proceed with the installation of a microwave relay system from Halifax to Saint John. This system was to employ time-division multiplexing in the 1700–1990-megacycle-per-second band. Two completely separate radio-frequency equipments were to be supplied as well as two separate multiplex equipments each capable of providing 23 voice channels. Only one multiplex equipment was to be provided with channel units. Subsequently, the second multiplex equipment was provided with channel units bringing the total circuit capacity to 46. In December 1954, a third radio-frequency equipment was added as a standby.

The final repeater sites were chosen from maps and ground surveys as shown in Figure 1. In choosing these sites, specific attention was paid to accessibility since this can be a problem in winter. Typical winter road conditions are shown in Figure 2. With the exception of Upper Sackville, it was possible to locate all Nova Scotia sites near an established roadway from which snow is regularly removed. Upper Sackville is on a maintained bush road half a mile (0.8 kilometre) from a highway. All Nova Scotia repeater points are within 5 miles (8 kilometres) of a major highway. In New Brunswick, a half-mile access road up Bloomsbury Mountain had to be constructed. All sites were within easy reach of existing power lines.

Figure 2—Typical conditions on a secondary road in Nova Scotia during winter.

Subsequent to the establishment of the Bloomsbury Mountain repeater, a new airport was constructed nearby, and the antenna tower was ruled an air navigational hazard. Consequently, a new location at Otter Lake was chosen, which although more remote actually provided a better path into Saint John. An access road approximately a mile long had to be constructed. The equipment was moved to this new site in the autumn of 1954, coincident with the installation of the third radio-frequency unit.

The tower heights were computed to yield at least 70 per cent of the first-Fresnel-zone clearance, using the true radius of the earth. This criterion is quite conservative and in keeping with the reliability required for toll telephone circuits. The resulting profiles are shown in Figure 3. Since the system has been in service, failures or noisy conditions due to propagation have been negligible.

The transmission lines and antenna gains were chosen so that the net attenuation of any given path would not exceed 75 decibels. This would provide a fading margin of 27 decibels on any one path at one time for a resultant (*F1A*-weighted) overall signal-to-noise ratio of 48 decibels. At this ratio, toll circuits must be removed from service. Alternatively, combined fades could occur on more than one path in such a manner that the overall ratio dropped by 19 decibels.

In planning antenna gains for the third radio-frequency equipment, it was decided to use only 10-foot. (3.1-metre) paraboloids. Experience with the first two systems indicated that the increased signal strengths would allow the use of vacuum tubes with lower transconductances in the receiver intermediate-frequency amplifiers. This can result in a considerable saving in vacuum-tube replacements that will eventually offset the additional cost of the larger antennas.

Automatic transfer was not provided for the radio systems. Instead, all connections between the multiplex (pulse modulator and demodulator) and radio equipments at terminals are through coaxial patch cords. The spare radio system is put in service simply by changing the patch cords at each terminal. Parallel jacks are provided so that this can be done with no interruption to service. The unused radio system is normally bridged across an operative system through a special video amplifier to ensure that it will be continuously available for service.

Patch cords and jacks are provided at repeaters also. This allows service-channel equipment to be inserted in any radio system and also

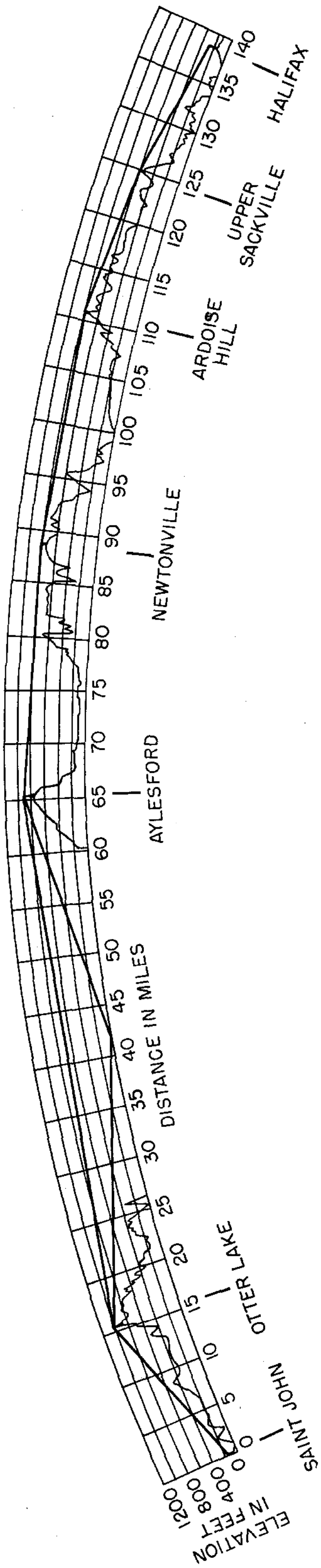

Figure 3—Profile of terrain along the route between Saint John and Halifax taken by the microwave relay system. The curvature of the earth is plotted to 4/3 actual radius.

provides means for changing radio systems at any repeater.

4. Basic Equipment

The type-*10B* pulse-time-modulated system of Federal Telecommunication Laboratories shown in Figure 4 was installed. The basic design of this equipment has been described elsewhere[1–3] and will be treated only briefly herein.

The channelling equipment provides multiplexing by time division. A train consisting of 23 channel pulses and one synchronizing pulse is developed. Each channel pulse is of 0.35-microsecond duration at the half-amplitude points and has a rise time of 0.15 microsecond. The synchronizing pulse actually is two such pulses spaced 1.3 microseconds from leading edge to leading edge. The spacing between pulses is 5.2 microseconds; so the length of one frame is 125 microseconds, and the frame-repetition rate is 8000 per second. Each voice circuit of the 23 that the equipment accommodates is assigned to one of the channel pulses and samples it once each 125 microseconds. The information is imposed on the pulse by moving it back and forth in time relative to the synchronizing pulse. This commonly corresponds to a deviation of ±1 microsecond. Therefore, the noise improvement is equal to the ratio of the pulse-deviation time to the pulse-rise time, or 16 decibels.

Separation of channels is accomplished at the receiving end by comparing the average position of each pulse with respect to the synchronizing pulse. Demodulation results from an exact comparison of each pulse position in a channel series with that of the synchronizing pulse.

Each of the 23 voice channels has a bandwidth of 200 to 3400 cycles per second and the total harmonic and spurious distortions do not exceed 5 per cent at 100-per-cent modulation. The bandwidth required to transmit the composite pulse

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[1] D. D. Grieg and A. M. Levine, "Pulse-Time-Modulated Multiplex Radio Relay System—Terminal Equipment," *Electrical Communication,* volume 23, pages 159–178; June, 1946.

[2] D. D. Grieg and H. Gallay, "Pulse-Time-Modulated Multiplex Radio Relay System—Radio-Frequency Equipment," *Electrical Communication,* volume 24, pages 141–158; June, 1947.

[3] S. Metzger, N. H. Gottfried, and R. W. Hughes, "Microwave Radio Links of Bonneville Power Administration," *Electrical Communication,* volume 29, pages 87–92; June, 1952.

train without pulse deterioration and consequent interchannel crosstalk is 2.8 megacycles per second.

The radio-frequency transmitter is a simple pulsed oscillator operating into a high-Q cavity with a frequency stability after warm-up of ±0.05 per cent. The oscillator operates in the frequency range from 1700 to 2100 megacycles per second. The composite pulse train from the multiplex modulator equipment is amplified and pulses the oscillator directly. A low-power continuous-wave oscillator, operating at the carrier frequency, is used as a "catalyzer" to stabilize the point on the leading edge of the modulating pulse at which the transmitter oscillates. This substantially reduces the random modulation that would be added by the transmitter. The transmitter delivers a peak power of 40 watts.

The radio-frequency receiver employs a continuous-wave local oscillator operating into a high-Q cavity, a crystal mixer, and an intermediate-frequency amplifier with a centre frequency of 30 megacycles per second. The total receiver gain, including video gain, is 85 decibels. An improved receiver, incorporating an untuned microstrip mixer crystal holder and an additional 15 decibels of intermediate-frequency gain, was supplied as part of the third radio-frequency equipment. This receiver exhibited a noise figure of 12 decibels compared to a noise figure of 14 to 16 decibels for the older receiver. A repeater installation is shown in Figure 5.

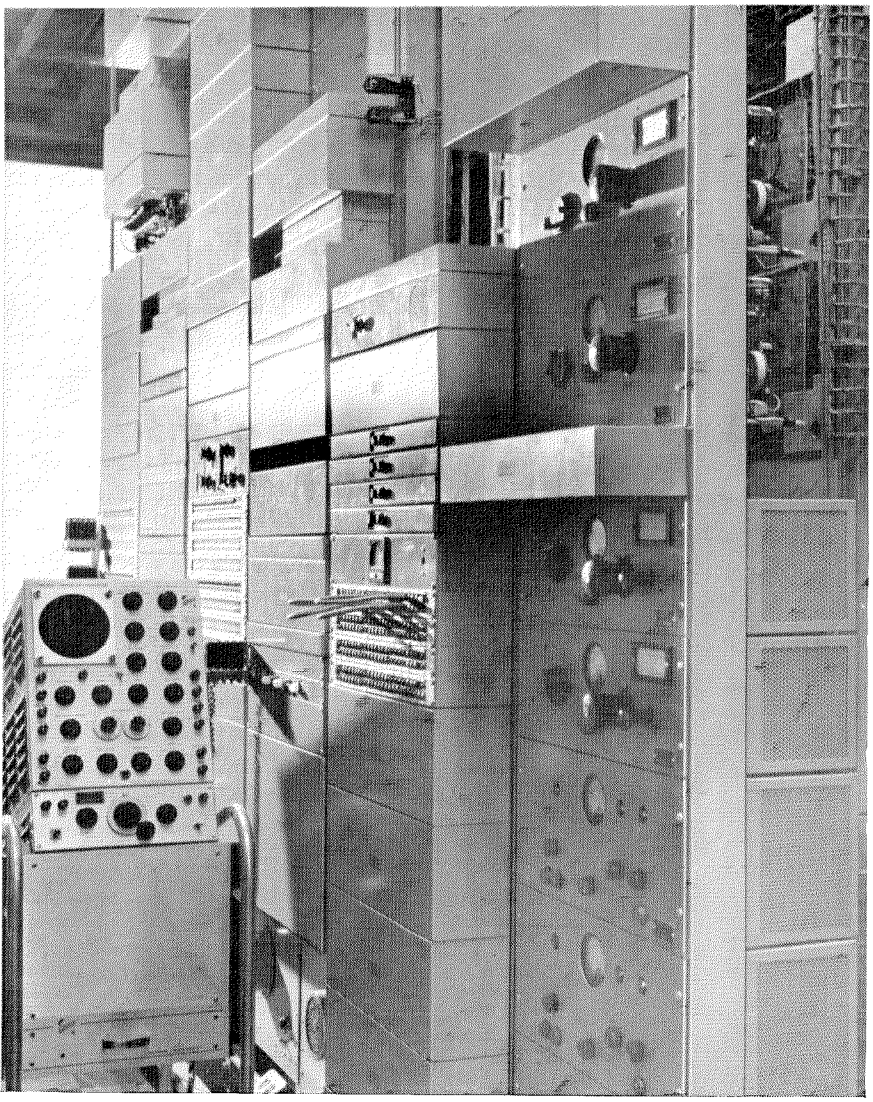

Figure 4—Terminal installation at Halifax. The radio-frequency equipment is in the foreground. Next is an alarm bay, beyond which are 4 bays incorporating 2 sets of multiplex modulator and demodulator apparatus. A test oscilloscope is on its trolley.

One innovation incorporated in this system was the demodulation of the information from the trailing edge of the pulse rather than the leading edge. This was required by the type of transmitter used. At a repeater, the transmitters operating in opposite directions are assigned the same carrier frequency, as is common practice in such systems. It was found that one transmitter tended to catalyze the other, increasing the random modulation noise on the leading edges of the pulses. This was eliminated by suppressing the leading edges and detecting the trailing edges. An additional advantage of trailing-edge detection is that the transmitter catalyzers become relatively insensitive. If the isolation between leading and trailing edges of the pulses were better, or if a 4-decibel degradation in signal-to-noise ratio could be tolerated, the catalyzer could be dispensed with completely.

5. Antenna Systems and Transmission Lines

With a pulsed-oscillator type of transmitter that is subject to frequency pulling, it is not practicable to multiplex transmitters on one antenna. It is possible however to duplex a transmitter and receiver on one antenna using a filter of the non-band-pass type. This system was therefore adopted and only one antenna is required for each radio-frequency system for each direction of transmission. For the 3 parallel systems, this requires 3 antennas at terminals and 6 at repeaters not employing diversity reception.

As previously mentioned, the antennas for the first two radio-frequency systems were designed to yield net path attenuations of about 70 to 75 decibels. This called for 6-foot-diameter (1.83-metre) paraboloidal antennas on all hops except the Bay of Fundy path, which required 10-foot (3.1-metre) paraboloids. Both sizes of antennas are energized by dipole–reflector assemblies and exhibit gains of 29 and 34 decibels respectively above an isotropic radiator.

For the third radio-frequency system, all paraboloids were 10 feet (3.1 metres) in diameter. These were of expanded steel mesh. A coaxial-to-waveguide transformation and a horn were used to illuminate the paraboloids. This method produces a low standing-wave ratio over a band from 1700 to 2400 megacycles per second. Future expansion may be accommodated therefore by multiplexing additional transmitters and receivers onto these antennas provided master-oscillator–modulated-amplifier transmitters are

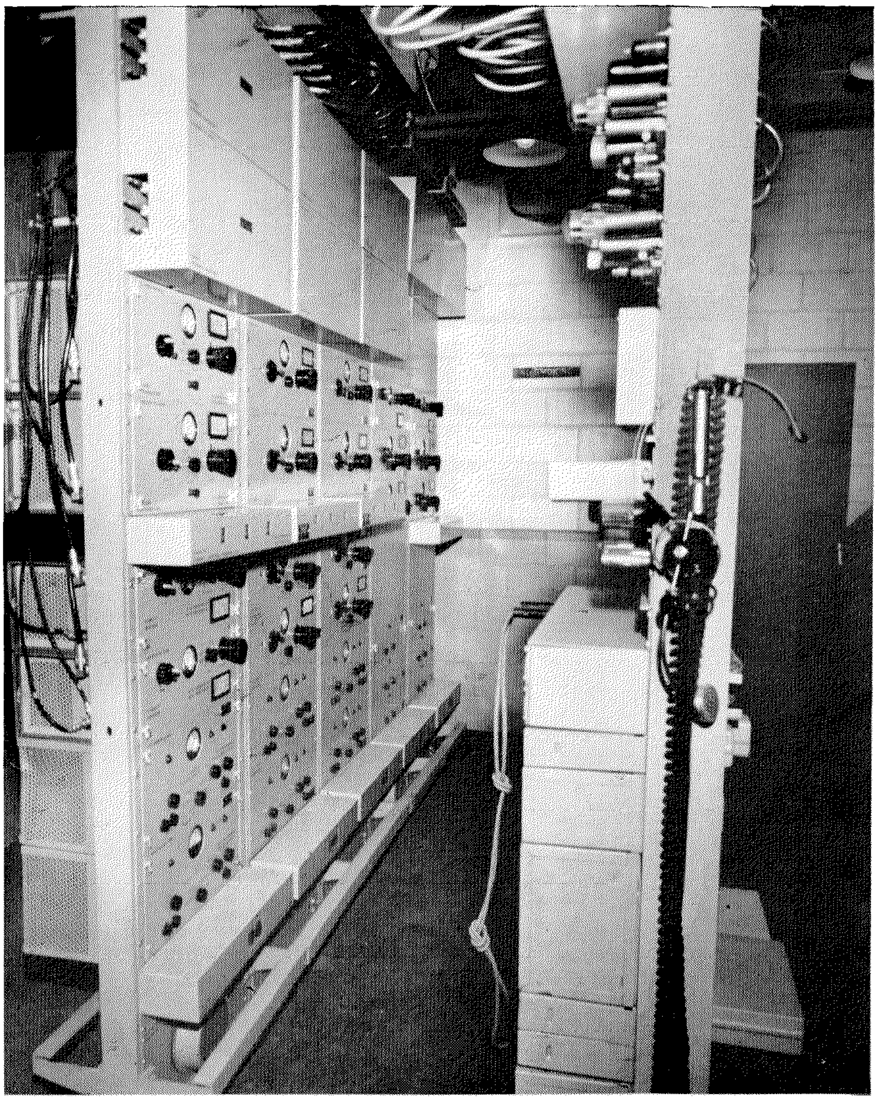

Figure 5—Repeater installation at Otter Lake. At the right are the alarm and the drop-and-insert equipment. The first 3 bays at the left are the radio transmitting apparatus and the other 2 bays are the diversity receivers.

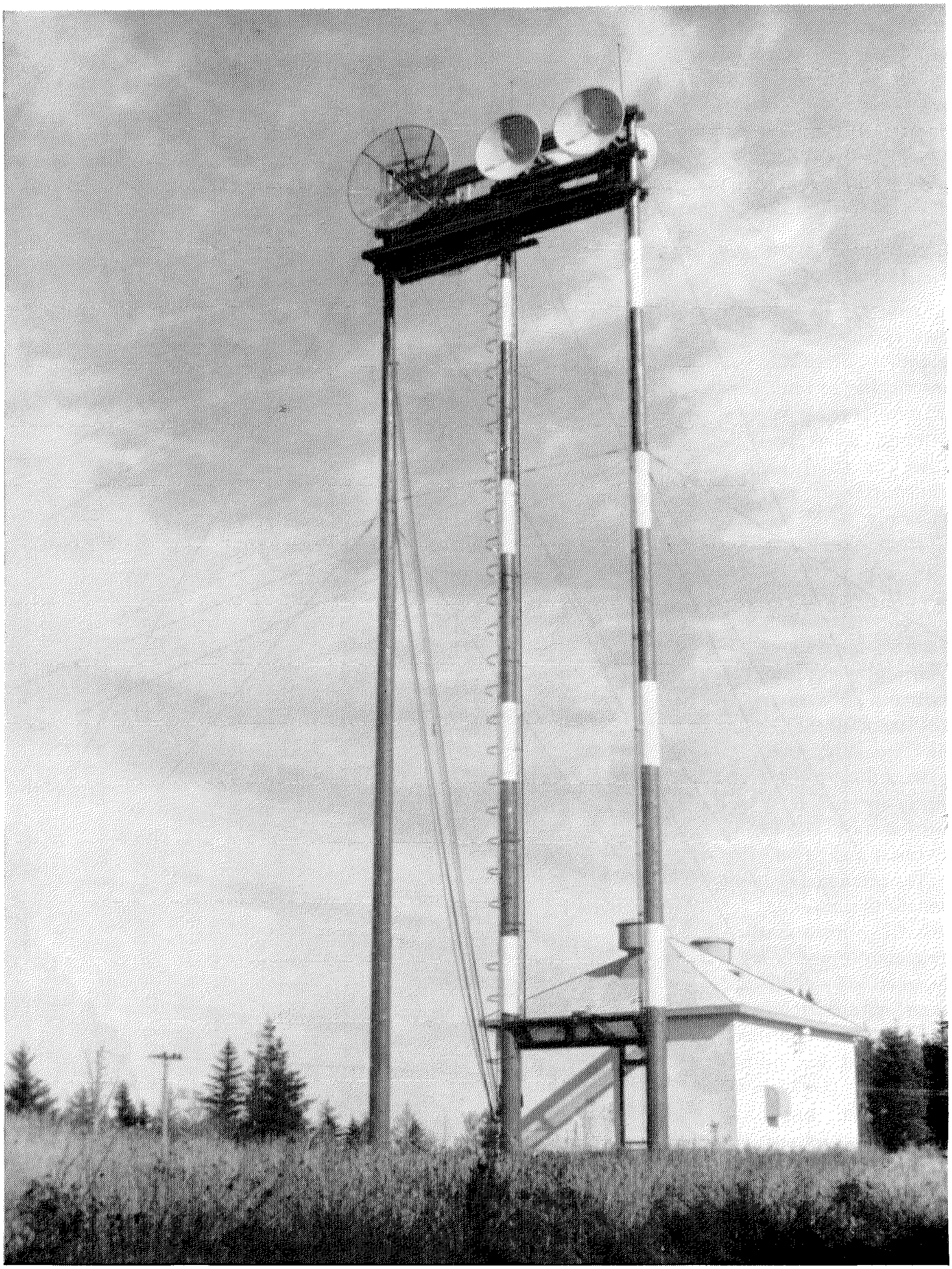

Figure 6—Repeater building and antenna arrangement at Upper Sackville.

used to permit multiplexing by means of simple band-pass filters.

At all repeaters except the two employing diversity reception, antenna elevations of 70 feet or less were required. It was possible therefore to use just two guyed poles with horizontal stringers between them to support the antennas. The four 6-foot (1.83-metre) antennas for the first and second radio systems at a repeater were supported between two poles. For the 10-foot (3.1-metre) antennas of the third system, another pole was added. The antenna arrangement at a typical repeater can be seen in Figure 6. This structure has been found very economical and possesses more than the required rigidity. It can be constructed entirely by a telephone or power company with a consequent saving in cost. The centre pole was equipped with a ladder and safety hoops for climbing. These hoops, which can be seen in Figure 6, have already prevented at least one serious injury.

At the repeaters at each end of the Bay of Fundy hop, a 127-foot (38.7-metre) tower was required for the diversity arrangement. A self-supporting steel tower was chosen to ensure the required rigidity. A view of the tower at Otter Lake is shown in the frontispiece; the one at Aylesford is essentially identical.

At each terminal, a steel billboard type of structure was erected on the roof of the telephone building. A problem arose at Halifax, where the 3-storey telephone building was designed so that a fourth storey could be added. To avoid interfering with such an addition, the steel framework of the building stair well was extended an additional storey and the billboard was mounted on top of this structure. This can be seen in Figure 7.

When the original radio systems were installed, ⅞-inch (22-millimetre) air-dielectric rigid transmission lines were used. By the time the third system was installed, semi-flexible line employing a helical insulator to support the centre conductor had become available and ⅞-inch (22-millimetre) cable of this type was used. At the same time, the complicated sections of rigid line employing three elbows to provide a flexible entry to each antenna of the first and second paths, were replaced with short lengths of the new cable. This improved the voltage standing-wave ratios considerably, reducing them to less than 1.5 to 1 throughout the band from 1700 to 2100 megacycles per second. This figure is more than adequate to prevent frequency pulling of the transmitter, which is the chief criterion in this type of system. For the same reasons, the entire runs at Saint John, which were particularly tortuous, were replaced with the semi-flexible cable. The continuously supported semi-flexible line has been found to be cheaper in initial cost, much easier to install, and requires essentially no maintenance.

Automatic pressurizing equipment was installed at all stations. Since the antennas are pressure-tight right up to the radiating element, the complete transmission-line system is protected from moisture. All transmission lines terminate on the top of the first radio-frequency bay and by means of coaxial patch cords any radio equipment can be connected to any antenna.

Semi-circular hoops were lagged to the antenna poles at the repeaters, and the transmission lines were mounted on the outside circumference of these. Since the hoops could hinge at the pole, expansion and contraction of the transmission lines were accommodated. At the two repeaters where steel towers were provided and at the terminals, the lines were fastened to the ladder safety hoops, which functioned in the same manner.

6. Bay of Fundy Diversity Arrangement

Based on the results of the propagation survey conducted over the Bay of Fundy, it was decided to use a three-receiver diversity system. Accordingly, at the Aylesford and Otter Lake repeaters, the following antenna arrangement was adopted. The antennas for the transmitters and main receivers of the three radio systems were located at the top of the 127-foot (38.7-metre) tower at an average height above ground of 120 feet (36.6 metres). The first diversity antenna was located at 60 feet (18.3 metres) above ground and a second diversity antenna 30 feet (9.1 metres) lower. Diversity receivers for each path were associated with each of these antennas. Multiplexing of the three receivers to each antenna was accomplished by band-pass filters.

A special diversity switching equipment was developed to select the receiver possessing the

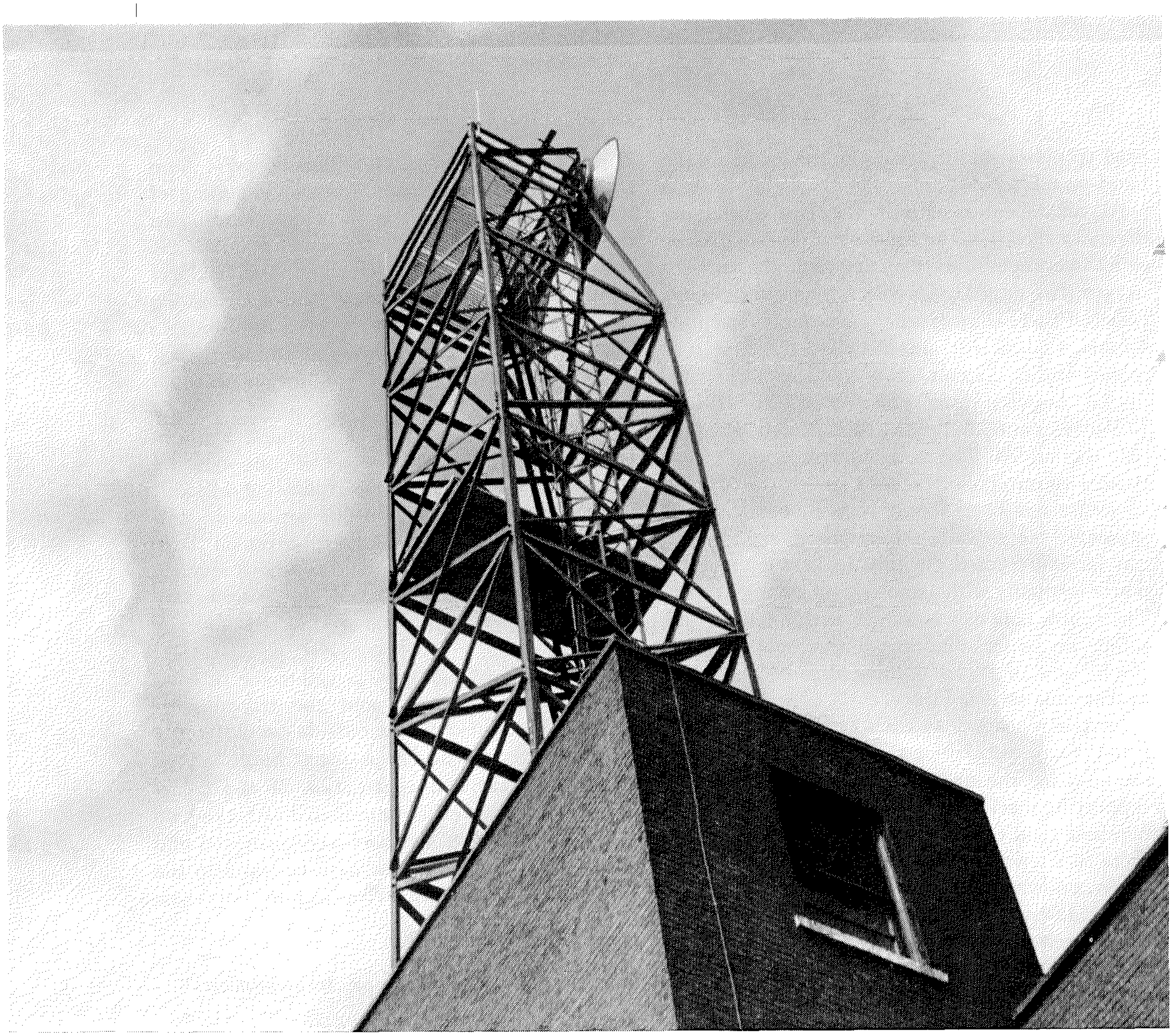

Figure 7—Halifax terminal showing how the steel structure of the stair well was extended to support the antennas and still permit another storey to be added to the building later.

highest signal. This unit monitors the automatic-gain-control voltages of the three receivers and biases off the video-frequency stages of the two receivers producing the weaker signals. This device operates on a differential of 0.3 volt, which corresponds to a signal-level difference of less than 2 decibels. To prevent the introduction of noise into the system by the switching process, the receiver being enabled is unbiased one second before the receiver being disabled is biased off.

While no specific data are presently available to establish the value of a three-antenna system over a two-antenna system, on several occasions when personnel happened to be at Aylesford or Otter Lake, it was observed that the main and normal diversity receivers were completely faded and the system was carried only on the remaining diversity antenna. It is true that compensation for changes in diversity spacing due to anomalous propagation can be obtained by choosing an antenna spacing less than the theoretical optimum. However, it is considered that in this case, in view of the reliability and overall signal-to-noise ratio required and obtained, the additional expense of a three-antenna system has been justified.

7. *Alarm System*

In a system intended to provide extremely reliable operation, it is important that all abnormal conditions throughout the link be automatically reported to control points. Means for doing this on radio relay systems for telephone companies may be simplified where local telephone lines interconnect the repeaters and terminals. In this system, it was possible to extend all alarms via wire lines to the terminals. The alarms are therefore transmitted regardless of whether the microwave link is operative or not. Continuity of operation of the alarm system is further ensured by using storage-battery power supplies. The alarms from the four Nova Scotia repeaters, Upper Sackville, Ardoise Hill, Newtonville, and Aylesford, are extended to Halifax. The alarms from Otter Lake are extended to Saint John. In addition, the primary alarms that indicate the status of the radio equipments at Otter Lake and at Saint John are extended via an existing land-line telegraph circuit to Halifax. The Halifax terminal, which acts as control for fault clearance, is thus cognizant of the operation of the radio equipment at all stations.

After considering the particular requirements of the alarm system for this link, it was felt that no suitable one existed, and therefore a new equipment was developed. An alarm coder at each repeater point is capable of sending a slow-speed pulse train that consists of a 1.5-second-long synchronizing pulse followed by 18 pulses occurring normally at a 4-pulse-per-second rate. Each of these pulses corresponds to a condition to be monitored. Should a particular monitored condition show a fault, the corresponding pulse is increased to 0.5 second from its normal 0.125-second duration. The message is therefore indicated by pulse width. On occurrence or clearance of any fault, the equipment sends one complete pulse train.

At each control point (Halifax and Saint John) one decoding equipment is supplied for each associated coding equipment. The decoder is prepared for operation and all alarms are cleared by the synchronizing pulses. The following alarm-condition pulses are then demodulated and those that are wide, corresponding to a fault, cause the appropriate alarm circuits to be energized. The alarms are actually displayed by means of switchboard lamps mounted in a standard jack field. The display board and decoding equipment at Halifax can be seen in Figure 4.

The capacity of this system can be increased in blocks of 18 alarm conditions by the addition of slave units to each coder and decoder. In addition to the alarm display, any alarm lights the aisle pilot and sounds the office audible alarm. A novel circuit is used to permit manual cut off of the audible alarm without preventing it from operating again if an additional fault is registered. Both remote and local alarms are tied into this system.

The alarms are transmitted from the repeaters to the terminals by means of frequency-modulated voice-frequency telegraph equipment. The highest-frequency channels are used in a speech-plus-duplex arrangement. In this way, the same line provides an order-wire channel.

The following alarms are transmitted from each repeater.

A. System *1*, west-to-east failure

B. System *1*, east-to-west failure

C. System *2*, west-to-east failure

D. System *2*, east-to-west failure

E. System *3*, west-to-east failure

F. System *3*, east-to-west failure

G. Commercial power failure

H. Diesel failure to start

I. Low voltage of 130- or 24-volt batteries

J. Low transmission-line pressure

K. Illegal building entry.

Four alarms are reserved for two future radio systems leaving three alarms for miscellaneous uses.

8. Service Channels

Three service channels are provided. One utilizes the alarm-system land line and is intended for use during emergencies, such as complete failure of the radio system. The other two are provided individually by the two multiplex systems. A drop-and-insert equipment is supplied at each repeater to provide access to either of these service channels. A drop-and-insert equipment removes the service-channel pulse from the pulse train and demodulates it. A new pulse, which may be modulated from a local telephone, is formed and is inserted in the pulse train in place of the removed one. The local telephone set and voice inputs and outputs of the two drop-and-insert equipments for the two directions of transmission are interconnected by a 12-wire hybrid as shown in Figure 8. This hybrid circuit isolates the two directions of transmission yet provides a 4-wire party-line connection for a local telephone. Amplifiers in the telephone send-and-receive leads serve to improve the matching and provide the required gain. By this arrangement, the service channel may be used as a party-line connection to all stations and provides at each point a circuit with a 6-decibel net loss to any other point. At any repeater, information is passed straight through the drop-and-insert and 12-wire hybrid without any net attenuation.

Alternatively, by manipulating two keys the service channel can be broken between any two repeaters. Independent conversations can then proceed on either side of the break. These same keys allow the drop-and-insert equipment to be switched to monitoring only, in which case the pulse trains are allowed to pass through unaltered although the service channel is demodulated on a bridging basis. When the drop-and-insert equipment is in circuit, its output is continuously monitored, and it will be automatically by-passed should it fail. This prevents the loss of the traffic circuits due to drop-and-insert-equipment faults. Outbound signaling to the repeaters is accomplished on the drop-and-insert equipment as well as on the Nova Scotia land-line wire by means of an amplifier speaker. It is believed that this system is the simplest and most reliable form of selective signaling. Signaling from repeaters to terminals may be accomplished in the same way. As an added means, a ringer receiver responding to 20 pulses per second of 1000-cycle-per-second tone is equipped at each terminal and a suitable signaling whistle is supplied at each repeater. Signaling on the New Brunswick land-line order wire is by means of magneto telephones.

9. Standby Power Equipment

Since the system carries important toll circuits, some of which are arranged for operator toll dialing, it is essential that no interruptions occur due to commercial power failure. All repeaters therefore are equipped with a no-break standby power system. The terminals were already equipped with non-automatic standby power facilities and these were deemed satisfactory.

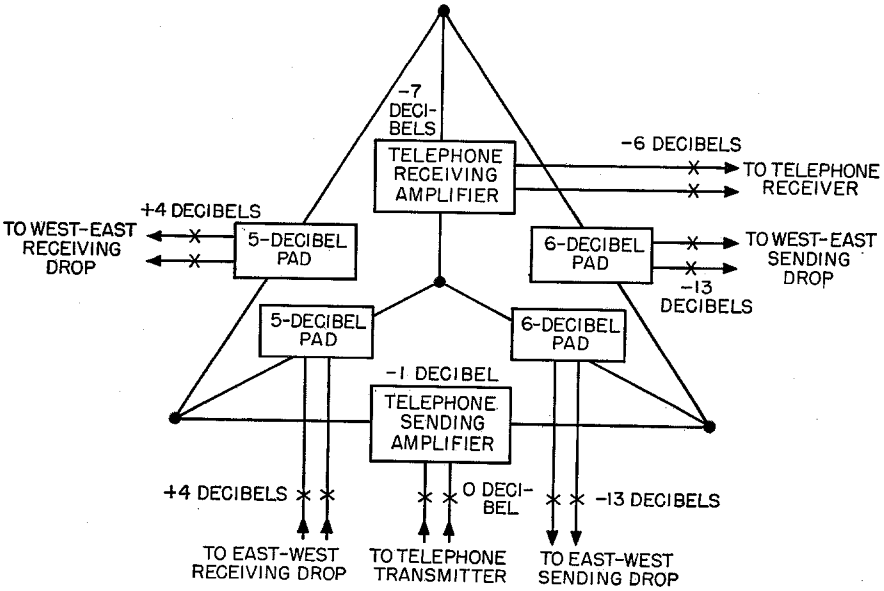

Figure 8—Block diagram of the 12-wire junction hybrid for a party-line service channel. The hybrid loss is 6 decibels. The levels indicated are in decibels related to 1 milliwatt.

In considering alternative designs for this system, it was decided that storage batteries offered the most reliable standby source. While they do require some service, telephone companies are very familiar with them and thus skilled in their maintenance. A well-maintained storage-battery plant can be depended on to supply power for a known period of time under practically any circumstance. Therefore a system employing a primary standby source derived from battery power was chosen. A secondary standby source was provided by a diesel-driven alternator. A diesel was chosen in preference to a gasoline engine due to its reliability and its relative ease of starting after lengthy idle periods.

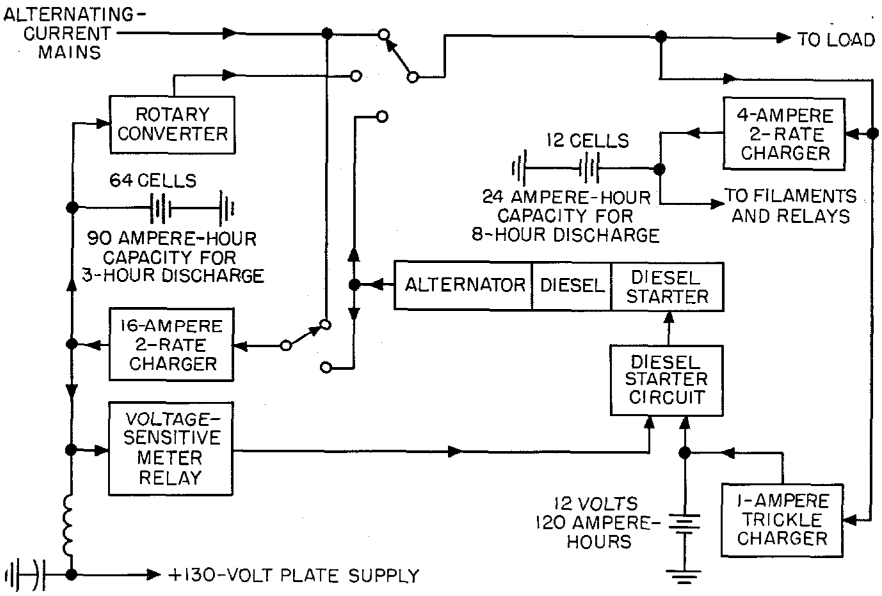

Figure 9—Power supply system with standby facilities for both short- and long-time mains power outages.

The system is shown in block-diagram form in Figure 9, and the various components can be seen in Figures 10 and 11. Power is normally supplied to the load directly from the alternating-current mains. The rotary converter runs idle and is driven from the direct-current side by the 130-volt battery. The running current for the converter is actually supplied by the 130-volt battery charger. This charger is a voltage-controlled two-rate type and maintains the battery in a fully charged condition.

If the mains supply fails completely or if the voltage drops below 95 volts, the load is automatically transferred to the alternating-current output of the rotary converter, which has been running at full speed and can deliver full power instantaneously. The rotary converter is governor regulated to maintain a constant output frequency of 60 cycles per second.

If the line power returns within a period of about 30 minutes, it automatically reassumes the load following a 5-minute wait. This period is introduced to prevent the line taking over on momentary returns, which often occur following a power failure. The 130-volt battery has a capacity of 90 ampere–hours for a 3-hour discharge rate and can carry the load for that period. When the terminal voltage of the battery drops to a predetermined value (normally corresponding to a 15-ampere–hour discharge or 30 minutes of operation), a contact-making voltmeter initiates the diesel starter circuit. This circuit is the three-strikes-and-out type; that is, three cranking cycles with a rest period after each one are provided. If after the three attempts the diesel has not started, the circuit is de-energized and an alarm initiated. To date, no cases of the diesel failing to start have been observed.

After a 1-minute diesel warm-up period, the normal load and the 130-volt battery charger are transferred to the alternator. The rotary converter is stopped to minimize battery drain. The alternator, in addition to supplying the load, charges the 130-volt battery at a 10-ampere rate. The alternator remains on load until the line power returns and until the battery plant is recharged to approximately 90 per cent of full capacity. In any case, the 5-minute delay previously described is introduced between the time when the line power returns and when the load is transferred to the line. After transfer, the diesel is stopped and the rotary converter is restarted.

A low-voltage relay initiates an alarm when the battery terminal voltage has dropped to 1.8 volts per cell. Another low-voltage relay operates at 1.7 volts per cell. This condition could occur only if the diesel had failed to start and the rotary converter was obliged to carry the load for 3 hours. In this case, the 1.7-volt-per-cell relay makes a final attempt to start the diesel and, if this fails, shuts down the entire plant to guard against battery damage.

A safety feature is provided that causes the diesel to start and run idle in preparation for receiving a load should the rotary converter fail. Alternatively this feature can be bypassed so that the diesel will not run continuously but will function on a regular standby basis. This type of operation would be used if extensive maintenance work were being done on the battery or rotary converter.

A further advantage of the 130-volt battery is its value as a plate supply for critical equipment such as the alarm system. A 24-volt battery was also included to act as a filament and relay supply for these critical equipments, to provide talking battery, and to energize the control relays of the standby power system. It has a 24-ampere–hour capacity for an 8-hour discharge rate. A 4-ampere two-rate battery charger is supplied to maintain the battery in a charged condition. This charger receives its power from the load circuit and is thus energized under all normal and standby conditions. A 1.8-volt-per-cell relay monitors the condition of the 24-volt plant in the same fashion as for the 130-volt plant.

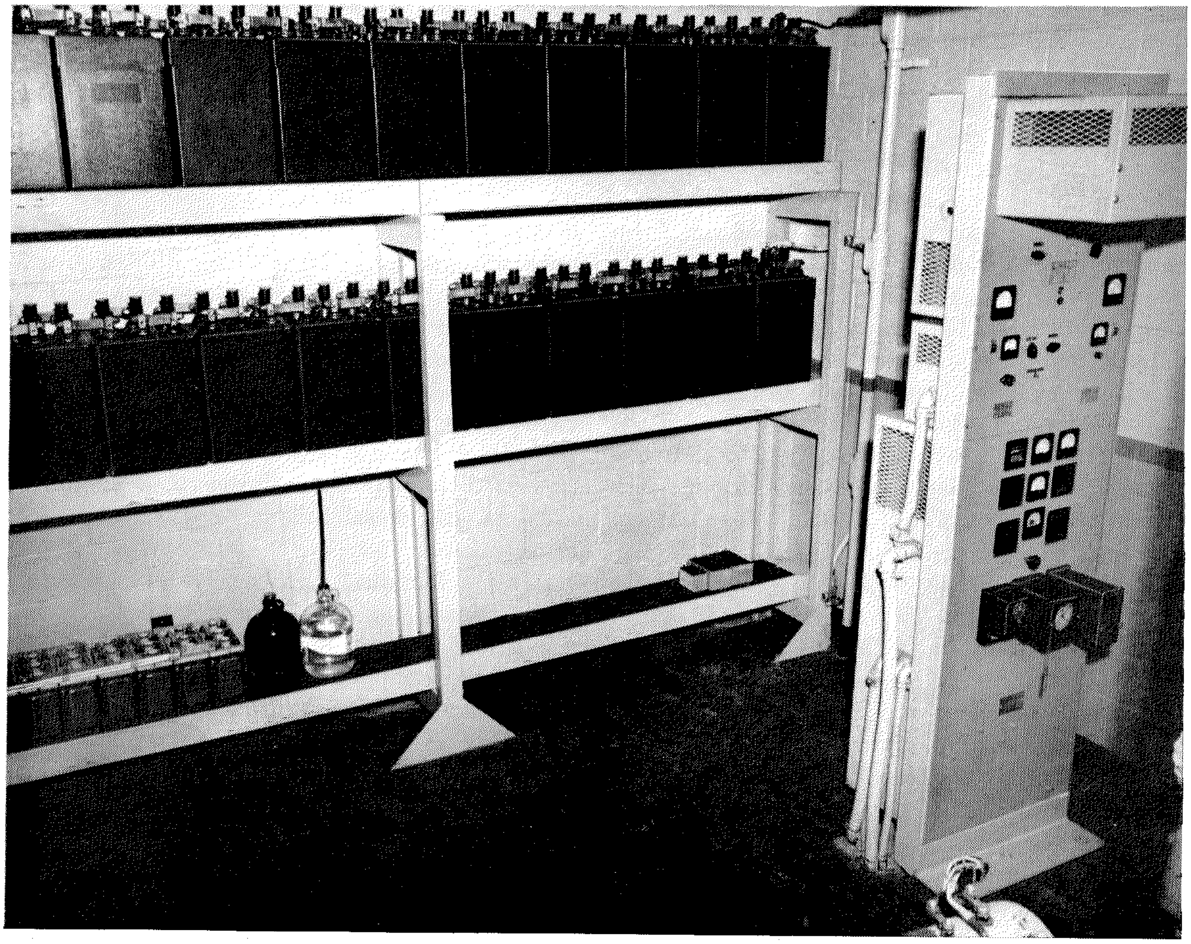

Figure 10—Power room at Otter Lake showing the 130- and 24-volt batteries, control panel, and battery chargers.

10. Test Equipment and Maintenance

In a pulse system, the principal test instrument is an oscilloscope. It permits the maintenance man to "see" the results of his work. A test oscilloscope having a special synchronizing attachment to allow any channel pulse to be selected and viewed was provided at each terminal. Another was supplied to the Otter Lake repeater primarily because of the difficulty of getting test equipment to that site during winter. A fourth oscilloscope was provided for use at the four Nova Scotia repeaters.

In addition to oscilloscopes, a microwave signal generator was supplied to each telephone company and a vacuum-tube voltmeter at each station. The telephone companies each had available a vacuum-tube checker and a noise-measuring set. A portable trailing-edge detector was supplied to the Maritime Telegraph and Telephone Company to allow accurate noise measurements to be made at repeaters. Lengths

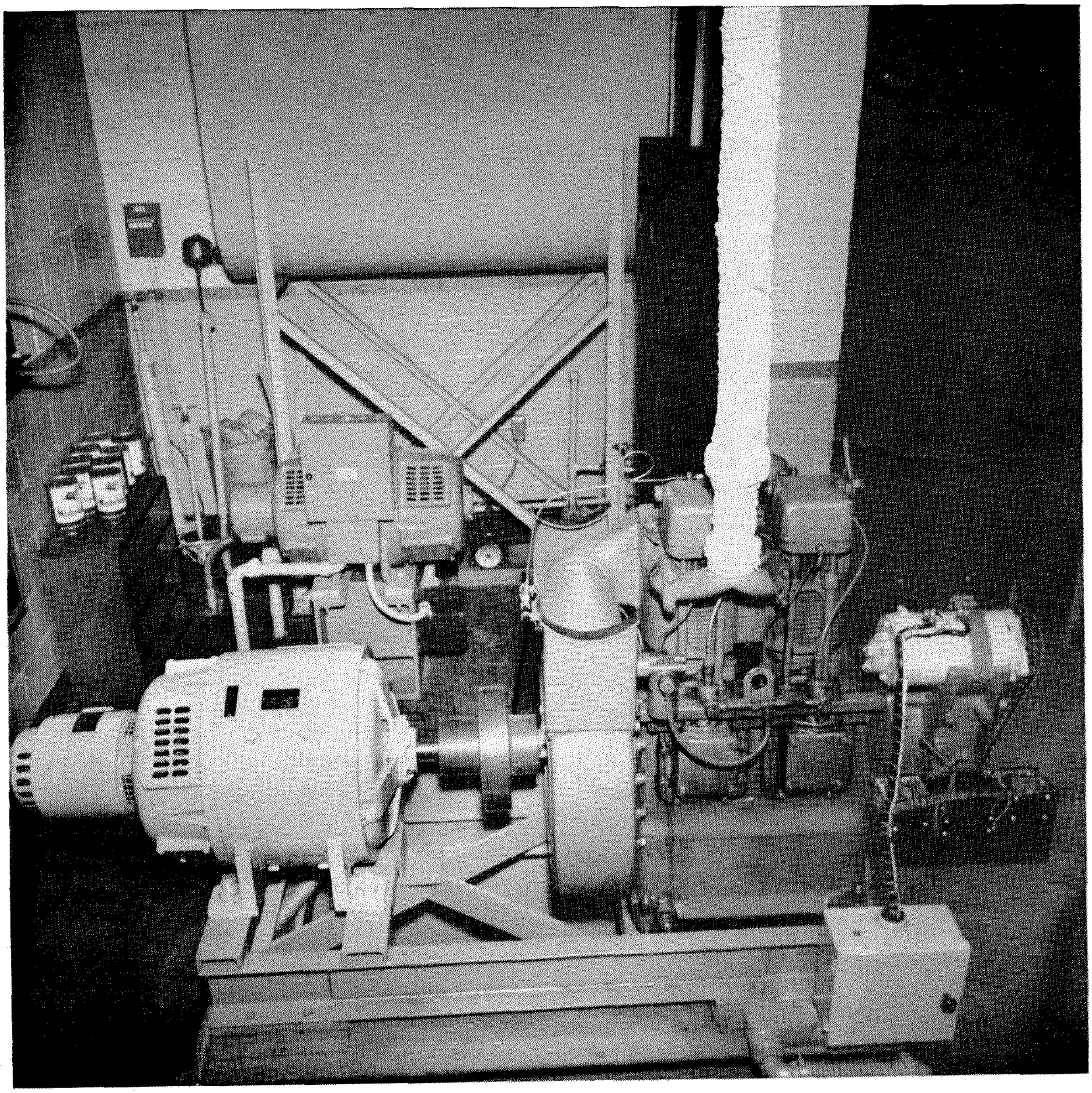

Figure 11—Diesel-alternator, rotary converter, and tank for diesel fuel at Otter Lake repeater station.

of lossy line producing 10, 20, and 30 decibels of attenuation were provided by the telephone companies to act as radio-frequency loads and to simulate fades. This equipment comprises the total test facilities.

Maintenance routines were based on the manufacturers technical knowledge of the equipment and the telephone companies general knowledge of toll maintenance. The experience gained during the initial stages of operation was of course taken into careful consideration. The routines thus developed have proved to be satisfactory. It is probable that further experience will alter the routines, most likely in the direction of reduced testing as a result of increasing skill of the service forces. The routines that were adopted can best be considered in two parts, multiplex-equipment maintenance and radio-equipment maintenance.

The terminal multiplex equipment is serviced completely by means of the test oscilloscope. The few key wave shapes in the common equipment are checked daily and the intermediate ones on a monthly routine. Vacuum-tube replace-

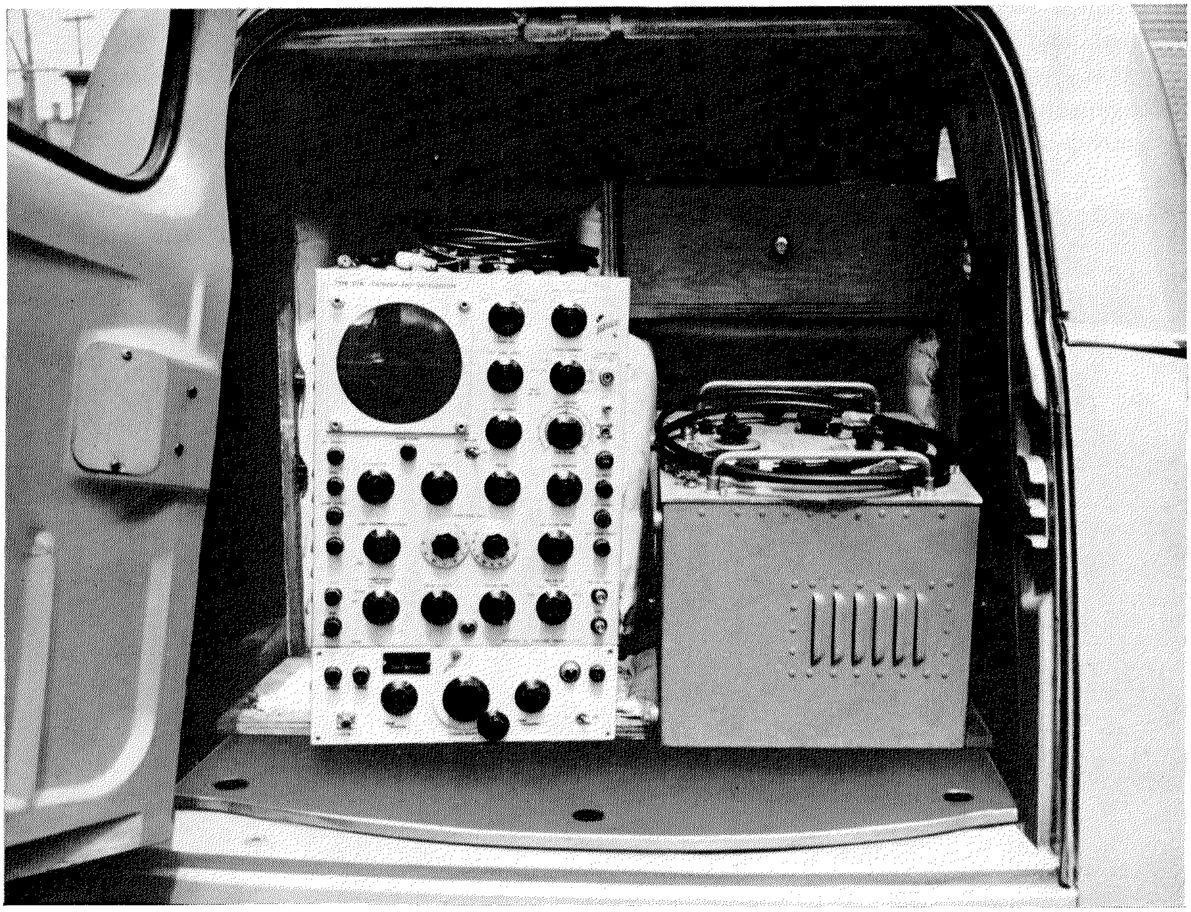

Figure 12—Rear view of maintenance truck displaying some of the test gear normally carried.

The Maritime Telegraph and Telephone Company acquired a panel truck for servicing the four Nova Scotia repeaters. This vehicle was fitted with padded compartments to accommodate the necessary test gear as shown in Figure 12. A microwave signal generator, vacuum-tube checker, oscilloscope, and portable trailing-edge detectors are normally carried in the maintenance vehicle.

Following the installation, a one-week instruction course was conducted for all personnel who would be associated with the system maintenance. This included both basic theory of the equipment and maintenance procedures.

ment is effected on the basis of these tests. In an amplitude-limited pulse system, a vacuum tube will operate beyond end-of-life transconductance values without appreciable change in output wave shape. At some point, deviation in output will then begin to occur quite rapidly. The waveform test routine described takes advantage of this characteristic to obtain maximum usage of the vacuum tubes without endangering equipment performance or reliability. The condition of the vacuum tubes in the electronically regulated plate supplies are checked by noting the range over which regulation is maintained.

The loss of a single channel is not considered serious, and therefore the vacuum tubes in channel units are run to destruction or until the channel circuit requirements cannot be met. The channel net loss is adjusted on a regular basis and daily operational checks carried out on each channel in exactly the same manner as for any toll circuit.

The radio-frequency equipment at terminals is checked once a week. The same schedule applies at repeaters if the roads permit access. Complete metering of the transmitter is provided including power output at normal and reduced filament voltage. The condition of the vacuum tubes can readily be ascertained from these readings and replacement made as required. The local-oscillator stage of the receiver is completely metered and these readings indicate the requirement for replacement of this vacuum tube. The receiver intermediate-frequency amplifier cannot be checked in this fashion. Therefore the receiver overall gain is checked once a month using a microwave signal generator. When the gain is reduced by a certain amount, any faulty intermediate-frequency-amplifier vacuum tubes are replaced using either a trial substitution method or the vacuum-tube checker as a criterion.

During the weekly visit to repeaters, the alarm system and the standby power system are given an operative check. The vacuum tubes (which are all 10 000-hour type) used in the alarm system are checked once each three months on the vacuum-tube checker. The service-channel equipment, including the drop-and-insert apparatus, is routine checked as required except that a preventative maintenance check is made once each 6 months. Antennas, antenna structures, transmission lines, and pressurizing equipment are carefully checked before and after each winter season.

11. Interference

In a microwave radio relay system, the signal-to-noise ratio is a function of the total noise power introduced into the system from four sources.

A. Multiplex noise; with this equipment this noise is comparatively small.

B. Transmitter noise; this is small if the transmitter is properly catalyzed.

C. Receiver noise; the effect of this is controlled by the receiver noise figure and the magnitude of the incident signal. As previously mentioned, the net path attenuations have been kept conservative, thus insuring that a large signal is applied to each receiver.

D. Stray sources.

In any radio relay system, noise power is always introduced into the system by stray paths. In this system, four separate stray sources introduced noise that far overshadowed the first three conventional noises discussed above. These stray sources were as follows.

A. Stray catalyzation of transmitters. As mentioned in section 4, it was found that transmitters physically adjacent and operating on the same frequency or adjacent frequencies catalyzed one another. Although the cause could be eliminated by using more frequencies or elaborate electrical screening, it was possible to eliminate the effect before demodulation by discarding the pulse leading edges and using only the relatively noise-free trailing edges. This incidentally places a lower limit on the pulse widths of the system because the pulse must have some flat top or the noisy leading edge will modulate the quiet trailing edge. The upper limit of pulse width is fixed by the duty-cycle capabilities of the system. To facilitate pulse-width maintenance, delay-line devices were installed ahead of the transmitters at the terminals and at the midpoint repeater in Newtonville. These automatically correct the pulse width to the optimum value.

B. Interference from transmitter to receiver at video frequency. This was caused by the high-level modulator pulse interfering with the second detector of an adjacent receiver. It was eliminated by simple shielding of the receiver.

C. Direct radiation from one antenna to another. This occurred where the antenna dipoles lay in parallel planes, that is, where horizontal polarization was in use and the antennas were mounted one above the other, or where vertical polarization was used and the antennas were mounted side by side. This effect was eliminated by inserting between the antennas suitable electric shields made of bonded-brass-wire cloth.

D. Signals arriving through the effective back-to-front ratio of an antenna. This subject has been studied thoroughly by others.[4] The theoretical back-to-front ratio of an antenna assumes it to be located in true free space. The practical or effective back-to-front ratio differs from the theoretical in that the antenna receives random signals reflected from surrounding terrain. Some of the possible cross-talk paths in a typical system are illustrated in Figure 13. The true front-to-back ratio of the antennas are of the order of 60 to 65 decibels. The unwanted signal will be at about the same level as the receiver noise and therefore its effect is not too serious. However, the actual interfering signals received on the third radio system, due to foreground scattering, were as high as 15 decibels above the noise and averaged about 10 decibels above the noise. With the receiver noise that exists and assuming a path attenuation of 70 decibels (a fair figure for the third radio system with 10-foot (3.1-metre) paraboloids), the interfering signals are about 53 decibels below the desired signal. This corresponds fairly well with results obtained on other systems. Unfortunately, it is impossible to eliminate this interference unless four frequencies are used for each radio system. This is impracticable as sufficient frequency assignments are not available. With a ratio of 53 decibels between the carrier and interfering signals and making allowance for the modulation improvement, weighting improvement, and cumulative effect of 6 tandem hops, an overall channel signal-to-noise ratio of 64 decibels (*F1A* weighted) would be expected. This corresponds exactly to the figure obtained.

This interference explains why the signal-to-noise ratio provided by the third radio system, which has 6 decibels more antenna gain per hop and a receiver noise figure better by 4 decibels than those of the first two systems, exceeds that of the first two systems by only 2 decibels and not 10 decibels. If this interference were eliminated by the use of four frequencies per hop, the radio signal-to-noise ratio would increase to 74 decibels, which combined with the multiplex noise would result in an overall channel signal-to-noise ratio of approximately 70 decibels.

There are two other types of interference worthy of mention. The first is overreach and the second is direct radiation from transmitter to receiver due to their proximity in frequency.

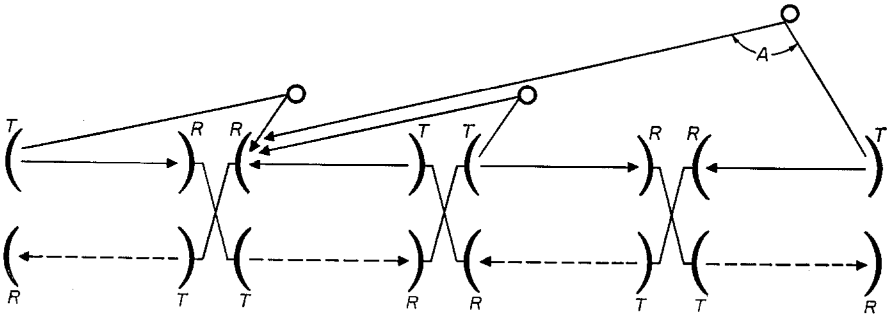

Figure 13—Some possible paths for cross-talk among the several radio links. Two radio frequencies are used, one being indicated by solid arrows and the other by broken-line arrows. *T* and *R* indicate the transmitting and receiving paraboloids. The terminal and repeater equipments are omitted. Reflecting regions are indicated by small circles. Path *A* is not necessarily a reflecting region: transmission can occur in this particular path through the normal methods of propagation (usually beyond the horizon).

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[4] H. W. Evans, "Cross-talk in Radio Systems Caused by Foreground Reflections," *Convention Record of the IRE 1953 National Convention*, Part 2, Antennas and Propagation, pp. 59–63.

Both of these were accounted for in the basic system engineering and did not occur. Overreach is radiation from one station to the third station away and can be effectively prevented by the geographical layout of stations. Direct radiation is dependent on the frequency assignments. In this system, the transmission frequencies of the parallel radio links are separated by 40 megacycles per second. The transmitter and receiver at any station are separated by 120 or 180 megacycles per second. This wide spacing is required since the transmitter and receiver are connected to a single antenna through duplexing filters. The closest spacing between a transmitter and a receiver of any system is 40 megacycles per second and the closest spacing between a receiver local oscillator and the carrier frequency of any other receiver is 10 megacycles per second.

12. Performance

There are three basic parameters in a time-division multiplex radio relay system that indicate the overall performance. These are the channel signal-to-noise ratio, the channel signal-to-cross-talk ratio, and the reliability. These three factors are functions of both equipment and system design. Other factors, such as channel frequency response and distortion, are solely dependent on equipment design. In installations employing wide-band pulse-position modulation, cross-talk is only slightly influenced by system design.

Table 1 shows typical interchannel cross-talk ratios. These figures were actually taken at Halifax with the modulator and demodulator equipments of one system patched back to back.

TABLE 1

SIGNAL-TO-CROSS-TALK RATIOS IN DECIBELS ON SYSTEM *1* AT HALIFAX

	Listening Channel																						
Modulated Channel	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23
1		74	72	70	62	73	74	77	69	72	68	74	67	75	75	72	71	79	66	74	70	74	69
2	80		79	80	74	80	80	74	80	80	76	80	80	75	80	80	75	80	80	76	80	77	78
3	78	78		74	74	72	80	80	72	80	80	72	80	80	71	80	80	72	80	80	69	80	80
4	74	80	80		78	80	73	80	80	75	80	80	72	80	80	74	80	80	74	80	80	69	80
5	80	80	80	80		80	80	75	80	80	80	80	80	80	80	80	80	80	80	76	80	80	80
6	78	76	76	78	76		75	74	70	80	80	80	76	80	80	80	79	80	80	80	76	80	80
7	80	75	75	73	75	75		76	73	72	80	80	80	80	80	80	79	80	76	80	76	64	79
8	80	79	78	74	76	76	76		78	75	80	80	76	76	80	80	80	80	80	75	80	77	76
9	80	80	74	80	80	71	80	80		75	80	80	80	80	75	80	80	79	80	80	75	80	80
10	80	80	80	80	80	80	74	80	74		80	72	74	80	80	80	80	80	80	80	80	66	80
11	80	80	74	74	78	73	75	75	73	77		66	74	76	80	80	80	78	79	80	76	78	80
12	80	80	80	80	80	75	80	80	77	80	80		73	78	72	80	80	76	80	80	80	80	80
13	79	80	80	75	80	80	75	80	80	79	80	80		75	73	71	80	80	80	80	80	67	80
14	80	70	80	80	70	80	80	70	80	80	70	80	80		70	80	68	80	80	80	80	75	74
15	80	80	75	80	80	71	80	80	73	80	80	75	79	80		80	80	70	80	80	75	75	80
16	80	80	78	80	80	80	80	76	74	80	78	78	75	77	71		75	72	69	80	80	75	76
17	77	77	78	80	76	78	80	80	74	80	80	78	74	77	75	75		80	73	80	74	75	75
18	80	80	79	80	80	80	80	80	76	78	78	80	75	75	75	75	74		69	76	68	78	80
19	80	80	80	80	80	80	80	80	80	80	80	80	75	79	80	78	77	78		78	79	65	79
20	80	80	80	80	80	80	80	80	80	80	80	80	77	78	80	78	78	73	78		80	71	79
21	80	80	80	80	80	80	80	80	80	80	80	80	79	80	75	80	78	79	73	76		69	75
22	80	80	80	80	80	80	80	80	80	80	75	75	80	80	77	75	77	80	80	80	80		73
23	76	68	79	78	76	78	80	95	75	75	73	73	73	72	75	75	80	80	80	70	73	75	

This method of checking multiplex cross-talk is preferred to operation over the radio path because cross-talk can be more easily distinguished from multiplex noise, which is lower than the radio system noise. In the table, the figure of 80 decibels represents the limit of the measuring equipment and actual ratios were often better.

Experience has shown that the inclusion of the radio path changes only the cross-talk from a channel to the immediately following channel (in time) and from the channels adjacent to the synchronizing pulse (channels 1 and 23) to all other channels. Table 2 shows these figures for the second multiplex system over the third radio path in the Saint John-to-Halifax direction. The channel signal-to-noise ratios are also shown. The noise in the first and second radio system is about 2 decibels higher. The minimum signal-to-noise ratio specified was 55 decibels for 99 per cent of the time. Thus safety margins of 5 decibels for the first and second systems and 7 decibels for the third system exist. With this margin, no difficulty has been experienced since the initial shakedown in maintaining the required signal-to-noise ratios.

The Maritime Telegraph and Telephone Company has kept a careful log of system outages. Tables 3, 4, and 5 summarize outage figures for 1953 through the first 5 months of 1955. In 1953, the first year of operation, total outage was 0.62 per cent or 0.09 per cent per station. In 1954, these figures were 0.69 and 0.097. It would be expected that outages would have decreased in 1954, which was the second year of operation, as

TABLE 2

SIGNAL-TO-NOISE AND SIGNAL-TO-CROSS-TALK RATIOS IN DECIBELS ON SECOND MULTIPLEX SYSTEM OVER THE THIRD RADIO PATH FROM SAINT JOHN TO HALIFAX

		Listening Channel																						
		1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23
S/N*		64	65	63	64	63	63	64	63	65	63	63	64	61	63	62	63	63	64	64	66	64	63	64
Modulated Channel	1		69	67	68	69	67	68	70	71	65	69	66	63	65	70	65	69	70	70	73	70	69	68
	2	68		61																				
	3		71		64																			
	4			69		69																		
	5				68		65																	
	6					69		64																
	7						69		69															
	8							70		71														
	9								69		67													
	10									71		67												
	11										69		68											
	12											69		65										
	13												68		65									
	14													67		68								
	15														67		65							
	16															67		69						
	17																68		68					
	18																	69		68				
	19																		69		70			
	20																			70		64		
	21																				70		67	
	22																					72		66
	23	61	70	65	64	67	65	66	67	69	65	65	66	69	65	66	67	65	66	66	68	64	65	

* Signal-to-noise ratio in decibels for *F1A* weighting.

a result of the clearance of initial troubles and the increased skill of the maintenance staff. In 1954 however, 35 circuits were in use, 23 on multiplex *1* and 12 on multiplex *2*. Since the third radio equipment had not been installed, there was no standby and any failure resulted in the loss of at least 12 circuits. These circuits were also lost while radio maintenance was in process. The advantage of the standby radio system can be clearly seen in the report for 1955, which shows a total outage figure of only 0.094 per cent or 0.014 per cent per station. It will be noted that in 1955 the fraction of the per-cent outage due to testing activities dropped to 5.42 per cent while the outages due to vacuum-tube failure rose to 64.41 per cent of the total. The balance among the various causes of outages during 1955 is considered as it should be. It will be noted that propagation outage in 1953 was 0.21 per cent of the total; in 1954, none; and in

TABLE 3

Monthly Summary of Outages in 1953

Based on 24-Hour-per-Day Operation and Both Scheduled and Non-Scheduled Outages

Month	Number of Circuit Troubles*	Circuit Hours in Trouble†	Per Cent Time in Trouble	Trouble Classification in Per Cent of Total Outage Time						
				Propagation	Power Mains	Tubes	Radio Equipment	Associated Equipment	Testing	Other
January	187	135.73	1.22	2.20	57.58	32.70	—	1.45	6.07	—
February	151	30.81	0.38	—	41.90	28.43	—	1.33	29.15	—
March	63	144.91	1.36	1.20	—	11.99	16.84	15.94	54.03	—
April	241	197.64	1.83	—	—	5.96	—	14.37	79.67	—
May	160	58.29	0.50	—	18.44	11.53	—	46.54	23.49	—
June	79	53.76	0.44	—	1.49	1.49	—	0.78	96.25	—
July	188	81.57	0.49	—	15.83	20.35	—	8.90	54.92	—
August	68	15.40	0.09	—	—	—	—	16.00	84.00	—
September	191	53.80	0.33	—	12.58	65.54	—	8.05	13.83	—
October	112	74.38	0.45	—	10.06	20.70	2.92	15.27	51.05	—
November	27	31.32	0.20	—	—	83.81	3.18	1.40	11.61	—
December	6	17.34	0.11	—	—	18.34	—	—	81.66	—
Total	1,473	894.95	7.40	3.40	155.88	300.84	22.94	130.03	585.73	—
Average	122.75	74.58	0.62	0.28	13.09	25.07	1.91	10.84	48.81	—

* Number of circuit troubles equals the number of troubles times the number of circuits affected.

† Circuit hours in trouble equals the outage time in hours times the number of circuits affected.

TABLE 4

Monthly Summary of Outages in 1954

Based on 24-Hour-per-Day Operation and Both Scheduled and Non-Scheduled Outages

Month	Number of Circuit Troubles	Circuit Hours in Trouble	Per Cent Time in Trouble	Trouble Classification in Per Cent of Total Outage Time						
				Propagation	Power Mains	Tubes	Radio Equipment	Associated Equipment	Testing	Other
January	149	351.05	1.79	—	0.57	3.17	0.38	25.10	68.37	2.41
February	187	169.53	1.09	—	1.63	21.30	—	61.15	15.92	—
March	174	58.68	0.34	—	—	1.55	1.96	72.12	24.37	—
April	2	1.93	0.01	—	—	—	—	88.08	11.92	—
May	92	7.50	0.05	—	—	39.87	—	—	55.60	4.53
June	154	11.88	0.07	—	—	51.90	—	5.50	18.50	24.00
July	132	40.32	0.19	—	3.50	62.40	—	21.80	12.30	—
August	272	398.24	1.78	—	16.90	6.50	26.50	0.40	47.60	2.10
September	423	167.63	0.78	—	8.27	5.88	3.23	13.18	66.67	2.77
October	171	150.63	0.67	—	—	33.76	—	3.68	61.01	1.55
November	567	260.87	1.07	—	7.74	41.80	1.74	20.40	19.10	9.22
December	312	113.38	0.42	—	—	1.27	—	18.15	75.46	5.12
Total	2,635	1,731.64	8.26	—	38.61	269.40	33.81	329.56	476.82	51.70
Average	219.58	144.30	0.69	—	3.22	22.45	2.82	27.46	39.75	4.31

1955, 2.48 per cent of the total. These figures, expressed as percentages of the total operating time, are only 0.0012, none, and 0.0023 per cent, respectively.

The allowable-outage figures specified were 1 per cent of the total (assuming 100-per-cent radio standby) of which 0.5 per cent could be due to propagation and 0.5 per cent to equipment failure. It can be seen that from the beginning the system has operated substantially within the overall limit, and after completion at the close of 1954 has been operating very well within the limit for equipment failures. Propagation outages have of course been negligible.

One point worthy of comment is storm performance. Since the system has been in operation, several severe storms involving snow, ice, and wind have been successfully weathered. For instance, when hurricane Edna swept Nova Scotia in 1954, causing extensive damage, all toll circuits out of Nova Scotia except those provided by the radio relay system were out of service. Although all Nova Scotia stations were operating throughout the storm on standby power, this system performed without break except for a very short period while the Halifax office was being transferred to standby power. Thus one of the main goals in choosing a radio relay system, that is, invulnerability to storms, has been attained.

13. Summary and Conclusion

The Saint John-to-Halifax radio relay system utilizes two 23-channel time-division-multiplex equipments and three microwave pulse-amplitude-modulated radio-frequency systems to provide 46 toll-quality voice circuits between Saint John in New Brunswick and Halifax in Nova Scotia via 5 repeaters. These circuits form an important part of the trans-Canada telephone toll network. This system has been in successful operation for two and one half years. As a result of this experience, the following conclusions may safely be drawn.

A. Time-division multiplex equipment of this type (broad-band pulse-position modulation) is appliciable to medium-density medium-haul requirements for toll-quality circuits.

B. Parallel systems may be employed to yield multiples of the basic group of 23 circuits.

C. When proper care is taken in system design, signal-to-noise ratios can be obtained that have sufficient margin above the requirements for toll circuits as not to cause trouble.

D. If path calculations and engineering are done on a conservative basis, outages due to propagation fading in the 2000-megacycle-per-second region can be practically eliminated.

E. The equipment can be maintained by telephone toll maintenance personnel so that outages due to equipment failure are reduced to a small figure (less than 0.05 per cent per station). This assumes that a standby radio system and a standby power source are provided.

TABLE 5

MONTHLY SUMMARY OF OUTAGES FOR JANUARY–MAY 1955

Based on 24-Hour-per-Day Operation and Both Scheduled and Non-Scheduled Outages

Month	Number of Circuit Troubles	Circuit Hours in Trouble	Per Cent Time in Trouble	Trouble Classification in Per Cent of Total Outage Time						
				Propagation	Power Mains	Tubes	Radio Equipment	Associated Equipment	Testing	Other
January	198	25.30	0.08	—	—	91.31	—	—	—	8.69
February	280	38.50	0.12	—	1.76	22.10	58.02	7.18	—	9.94
March	270	27.63	0.09	12.40	2.30	76.80	2.30	—	3.90	2.30
April	82	2.64	0.01	—	—	61.50	15.30	—	23.20	—
May	352	55.00	0.17	—	4.78	69.32	1.59	9.96	—	14.35
Total	1,182	149.07	0.47	12.40	9.84	321.03	77.21	17.14	27.10	35.28
Average	236.4	29.81	0.094	2.48	1.97	64.20	15.44	3.43	5.42	7.06

F. A 50-mile (80-kilometre) overwater path at 2000 megacycles per second is practical if a space-diversity arrangement is used. If the path is no worse than grazing under the poorest conditions of abnormal refraction, outages due to propagation fading will be small. Some improvement is apparently obtained through the use of 3 receiving antennas properly located, but the exact degree of improvement to be expected is not known.

This system has confirmed other known advantages of microwave radio relay systems. Some of these are: invulnerability to storms, low initial cost, concentration of maintenance at a few points, and shorter possible routes. Because of the ability of a microwave system to span the Bay of Fundy, the toll-circuit distance from Halifax to Saint John was greatly reduced. This resulted in better toll service at substantially reduced rates.

14. Acknowledgments

The author acknowledges with gratitude the efforts of the staff of Standard Telephones & Cables Manufacturing Company (Canada), Federal Telecommunication Laboratories, Federal Telephone and Radio Company, and Federal Electric Corporation, who assisted in the planning and installation of this system, and most especially acknowledges indebtedness to Federal Telecommunication Laboratories for its technical assistance and to the members of the engineering and toll test departments of the Maritime Telegraph and Telephone Company and the New Brunswich Telephone Company for their valuable technical advice, aid, and cooperation throughout the installation period. Without the combined efforts of all these groups, this system could not have reached so successful a conclusion.

Recent Telecommunication Development

"Transistors and Other Crystal Valves"

Mr. T. R. Scott, director of research of Standard Telecommunication Laboratories, is the author of a recently published book on "Transistors and Other Crystal Valves" issued by Macdonald and Evans, 8 John Street, Bedford Row, London WC1, England. The book is intended for engineers who may be required to make use of these devices. It is divided into 10 chapters, 2 appendixes, a bibliography of 277 items, a short list of abbreviations, and an index.

Chapter 1—Introduction
Chapter 2—Crystal Imperfections
Chapter 3—p-n Junctions
Chapter 4—Point-Contact Devices
Chapter 5—Operating Temperatures—Output —Life
Chapter 6—Applications—Circuitry—Testing
Chapter 7—Special H.F. Transistors
Chapter 8—Choice of Semiconductors
Chapter 9—Conclusions—The Future
Appendix A—Band Structure of Semiconductors
Appendix B—Testing Techniques

The book is $5\frac{1}{2}$ by $8\frac{3}{4}$ inches (14 by 22 centimeters) and is bound in hard covers. It contains 236 pages of text, 17 pages of bibliography, 1 page of abbreviations, 4 pages of index, and 16 pages of preliminary material. Its price is 45 shillings. Copies may be obtained in the United States from Essential Books, Incorporated, 114 Fifth Avenue, New York, New York, at $7.20 each, postpaid.

Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands[*]

By ROGER McSWEENY
American Cable and Radio Corporation; New York, New York

PUERTO RICO, together with Cuba, Hispaniola, and Jamaica, comprise the Greater Antilles or West Indies; Puerto Rico being the most easterly of the group. About 50 to 75 statute miles (80 to 120 kilometers) farther east of Puerto Rico are the Virgin Islands, which are at the end of an extensive chain of islands called the Lesser Antilles. This paper is concerned with Puerto Rico and with Saint Thomas and Saint Croix, two of the Virgin Islands, which are about 18 degrees north of the equator, 1000 miles (1600 kilometers) east of Miami, Florida, and 1600 miles (2600 kilometers) southeast of New York City. The Virgin Islands are closely associated through both travel and trade with Puerto Rico and, consequently, require intercommunication with it and through it to the American mainland. Puerto Rico is a commonwealth associated with the United States of America while Saint Thomas and Saint Croix are territories of the United States.

Prior to installation of the very-high-frequency links to be described in this paper, telecommunication facilities available for public use were limited to submarine telegraph cables and to high-frequency radiotelephone and radiotelegraph circuits. The latter were not very satisfactory because of unstable sky-wave transmission over distances of about 75 miles and the high levels of noise and interference on amplitude-modulated frequencies between 4 and 7 megacycles per second.

The new facilities operated by the Radio Corporation of Porto Rico and All America Cables and Radio are based on a 150-megacycle frequency-modulated system that provides several telephone channels that connect with the public telephone systems of the islands. In addition, telegraph printer channels are provided to cable offices for general public use and other printer channels are leased to airlines and shipping companies. The telephone services are extended to the mainland through high-frequency radio circuits from Puerto Rico to the United States. Telegraph services are extended to the mainland through the cable or radio facilities of All America Cables and Radio.

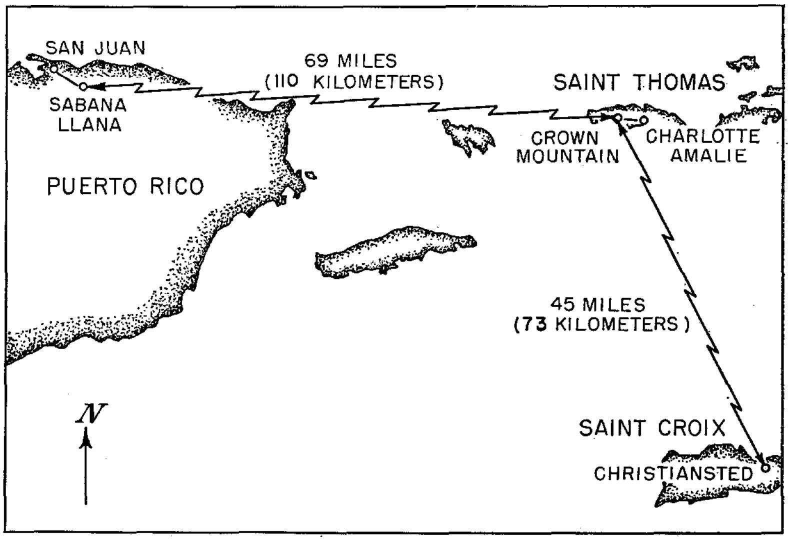

Figure 1—Map of radio system.

1. General Description

Figure 1 is a map of the system. There are two very-high-frequency radio links involved, one of 69 miles (111 kilometers) from Sabana Llana, Puerto Rico, to Crown Mountain on Saint Thomas and one of 45 miles (73 kilometers) from Crown Mountain to Christiansted on Saint Croix. At the Puerto Rican end, the Sabana Llana radio station is connected through 10 miles (16 kilometers) of cable to the telephone exchange and telegraph office in the city of San

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[*] Reprinted from *Communication and Electronics*, volume 74, pages 781–785; January, 1956. The illustrations are not identical with those in the original publication.

Juan. Both radio equipment and carrier apparatus for deriving the individual telephone channels are located at Sabana Llana.

On Saint Thomas, the Crown Mountain radio station is connected through 4 miles (6.4 kilometers) of cable to the town of Charlotte Amalie, where carrier equipment, telephone exchange, and telegraph terminals are located. On Saint Croix, the radio and terminating equipment are both at the telephone office in the town of Christiansted.

Voice and carrier provide 4 telephone channels on the radio link from Sabana Llana to Crown Mountain and this may be increased to 5 channels if required. Of the present 4 channels, the first is transferred at Crown Mountain to the Christiansted link to provide a direct telephone circuit between Saint Croix and San Juan. The second is connected at Sabana Llana to other radio equipment to provide a direct telephone circuit between Charlotte Amalie and New York. The third is for telephony between Charlotte Amalie and San Juan, and the fourth is subdivided by voice-frequency carrier equipment into 4 telegraph printer channels. This circuit may be increased to 9 printer channels, if needed.

Voice and carrier provide 3 telephone channels on the radio link from Crown Mountain to Christiansted and this may be increased to 4 channels, if required. Of the present 3 channels, the first is transferred at Crown Mountain to San Juan; the second provides a telephone circuit between Charlotte Amalie and Christiansted, and the third now carries a telegraph printer channel. This may be increased to 4 printer channels, if needed.

TABLE 1

Radio Frequencies

Frequency in Megacycles	From	To
158.01	Sabana Llana	Crown Mountain
152.75	Crown Mountain	Sabana Llana
152.57	Crown Mountain	Christiansted
157.83	Christiansted	Crown Mountain

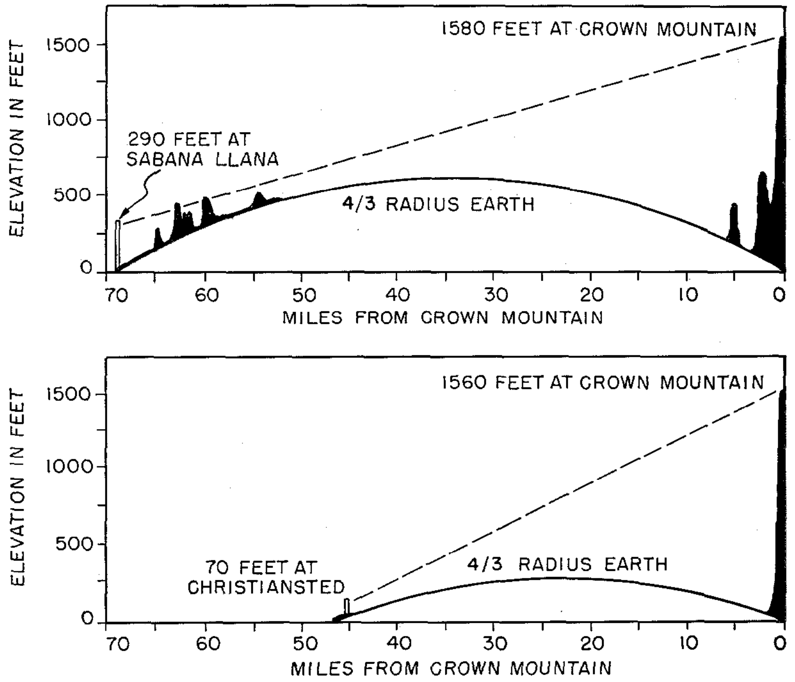

Figure 2—Profiles of radio paths.

The system may be expanded in the future to connect two other Virgin Islands, Saint John and Tortola, to Saint Thomas through the Crown Mountain radio station.

2. Frequencies and Sites

The radio frequencies employed are listed in Table 1. While lower frequencies in the very-high-frequency band probably would produce better results over these paths, the frequencies used are the lowest for which licenses could be obtained.

Prior to selecting the station site on Saint Thomas, a small transmitter was set up at Sabana Llana and operated for six weeks on 158.01 megacycles with a corner-reflector antenna mounted 260 feet (80 meters) above ground. Received field strength was observed at 19 places on Saint Thomas and continuous recordings for one week each were made at 6 locations. These tests indicated that a substantially stronger and steadier signal could be obtained at the top of Crown Mountain than elsewhere on the island.

As a result of this rather remote station location, the cost of radio equipment at Crown Mountain was only a fraction of the cost of necessary auxiliary facilities such as improvements to roadway, construction of power and telephone lines, building, et cetera.

3. Radio Paths

Figure 2 shows profiles of the two radio-frequency paths. The elevations shown are the heights of the antennas above mean sea level. Antenna elevation of 1580 feet (480 meters) at Crown Mountain is largely responsible for the satisfactory performance of the system. From this elevation, the 45-mile (73-kilometer) path to Christiansted is above the geometric horizon and presents little difficulty although a first Fresnel-zone clearance is not obtained. The 69-mile (111-kilometer) path to Sabana Llana falls below the "smooth-earth geometric horizon" but is above the "effective radio horizon" based on a $\frac{4}{3}$-radius earth. Tests indicated that the maximum practical height of the antenna (290 feet (99 meters) above sea level) using an existing tower at Sabana Llana produced signals that were 3 and 10 decibels better, respectively, than for heights of 120 and 70 feet (36 and 21 meters) at the same location. The maximum height does not clear the tops of two hills that project into the path as may be seen in Figure 2.

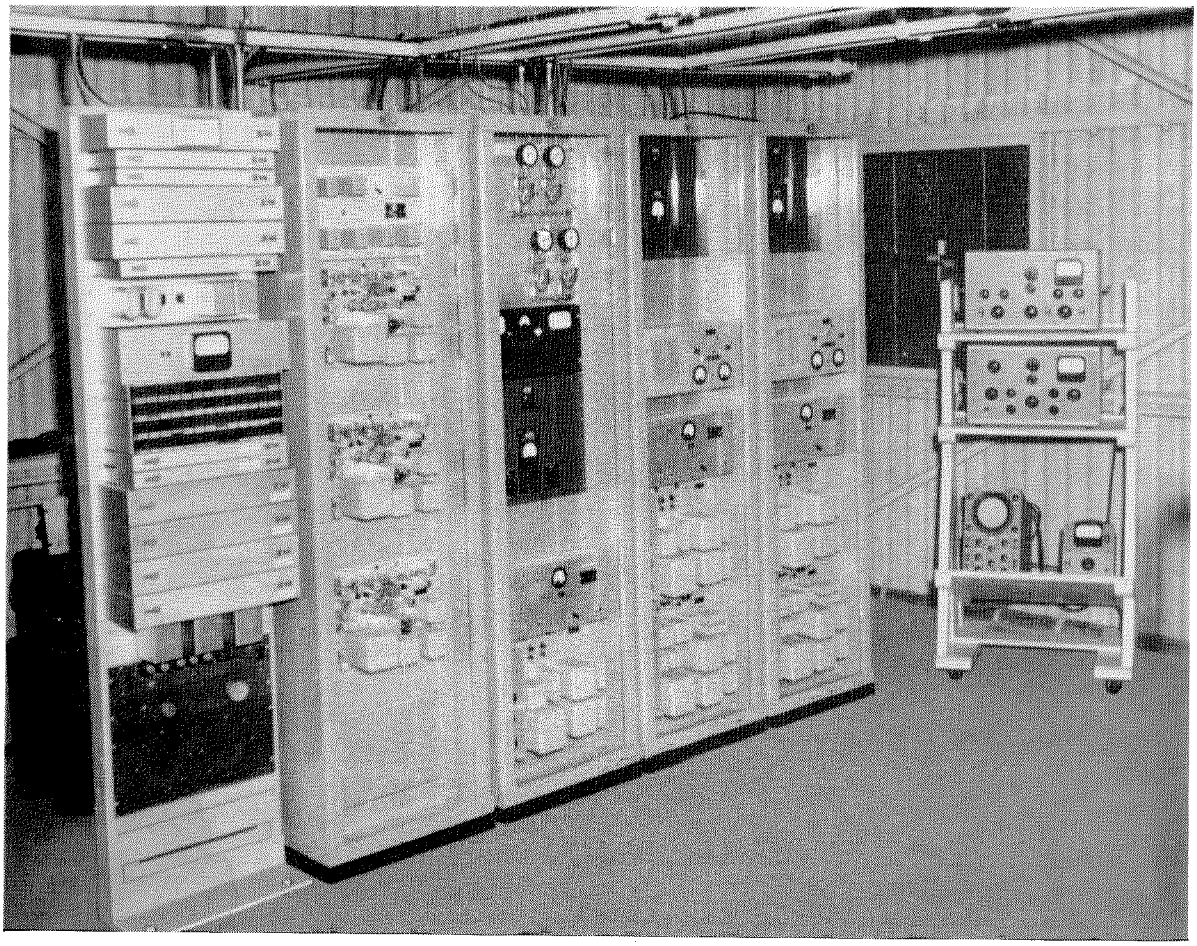

Figure 3—Installation at Crown Mountain. The left-hand bay is for line-terminating apparatus, the next rack holds three receivers, the center bay has coaxial-line pressure gages and valves and frequency monitors, while the remaining two bays are of transmitting equipment. Testing apparatus is in the small frame at the extreme right.

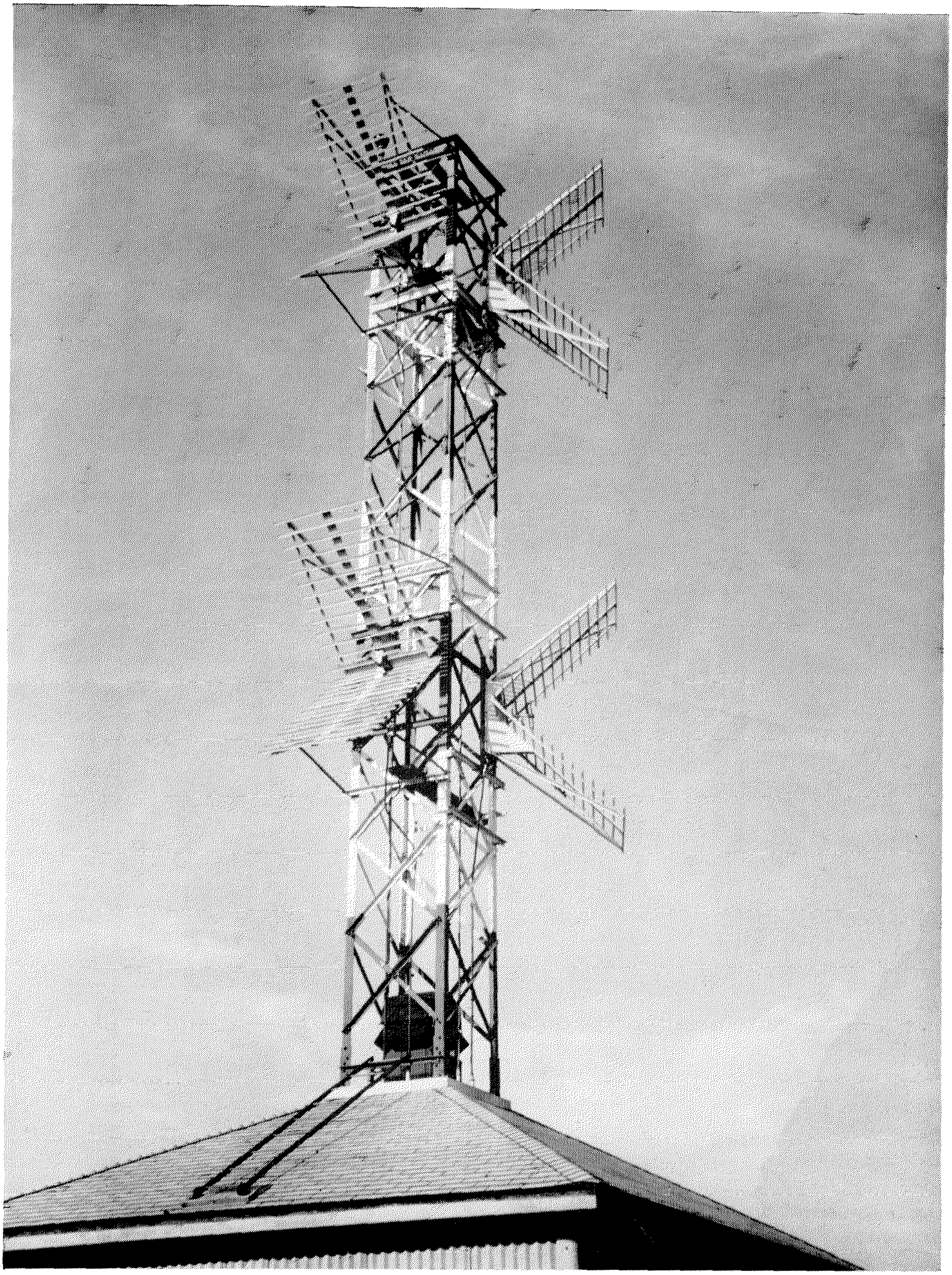

Figure 4—Original installation of 4 antennas at Crown Mountain.

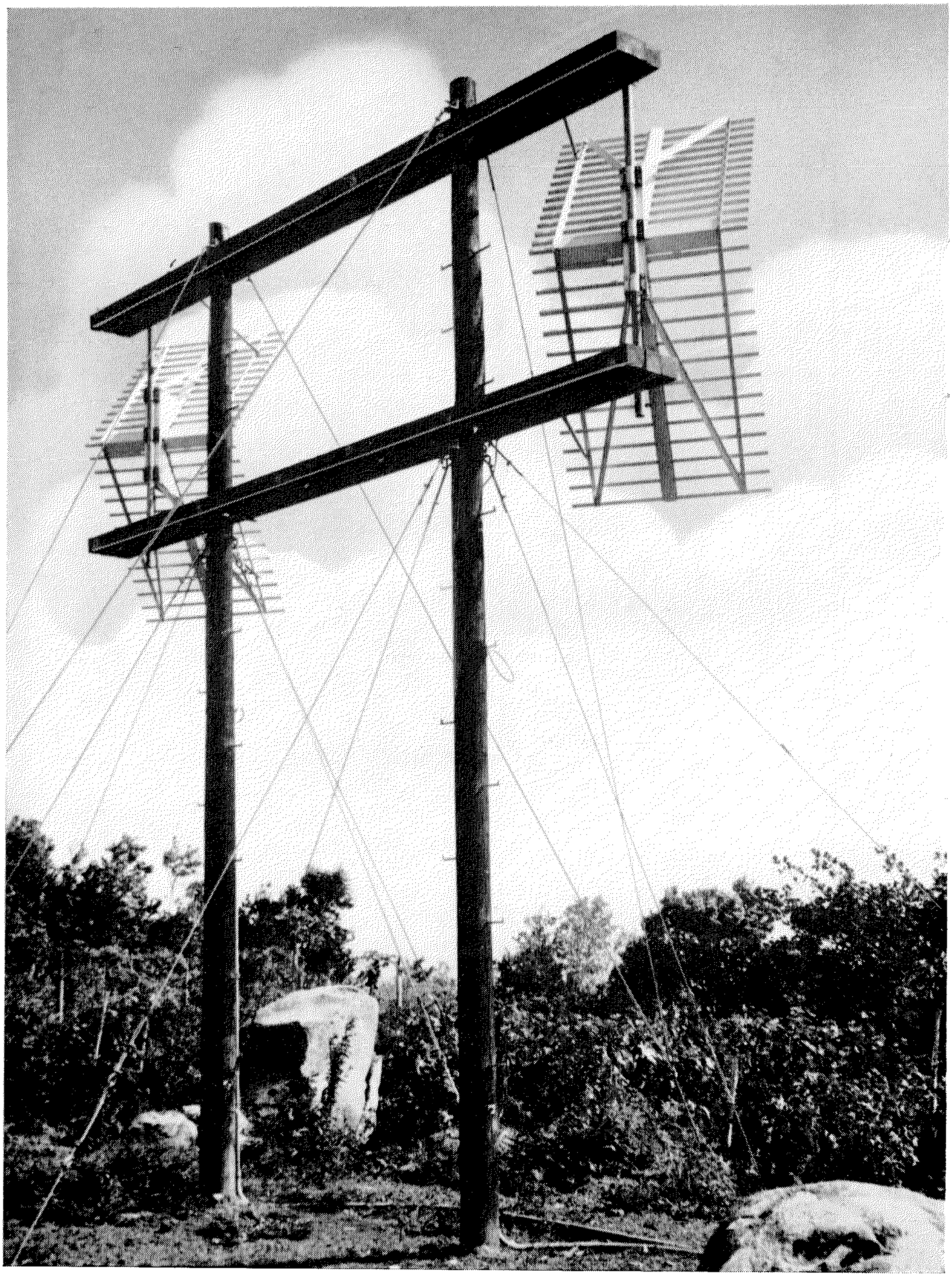

Figure 5—Second antenna installation at Crown Mountain.

Measured space losses (excluding estimated antenna gains and line losses) on the paths from Crown Mountain to Christiansted and Sabana Llana are opproximately 110 and 120 decibels, respectively. Calculated free-space losses for the same distances between isotropic antennas are 113 and 117 decibels. The low loss on the Christiansted path appears to indicate that signals reflected from the surface of the sea between the islands arrive approximately in phase with the directly propagated signals.

Usually there is little fading on either link, but slow fluctuations in field intensity of the order of 3 decibels occur continuously. In the first 18 months of commercial operation that started in August 1953, several periods lasting from 5 to 15 minutes have been observed during which field intensity dropped to approximately 15 decibels less than normal in both directions on the Sabana Llana–Crown Mountain link. Commercial operation continued without interruption during these periods. No correlation of the abnormally low fields with weather or time of day was observed.

4. Transmitters and Receivers

The frequency-modulated transmitters employed on the Christiansted–Crown Mountain link have a power output of 25 watts. Similar transmitters followed by power amplifiers of 100 watts output are employed on the Crown Mountain–Sabana Llana link. Peak deviation is limited by regulations of the Federal Communications Commission to ±15 kilocycles for this band. Limiting amplifiers are installed ahead of the transmitters to prevent excessive deviation due to unusual line levels. An outstanding feature of the transmitters is their low level of distortion.

Figure 3 shows typical transmitting and receiving equipment. The center cabinet contains air-pressure indicators for coaxial lines, a frequency and shift monitor, a limiting line amplifier, and a transmitter with power supply. The two cabinets at the right end each contain a limiting line amplifier, a transmitter with power supply, and a power amplifier with power supply.

The receivers are crystal-controlled double-conversion superheterodynes with automatic volume control, 2-stage limiting, and discriminator.

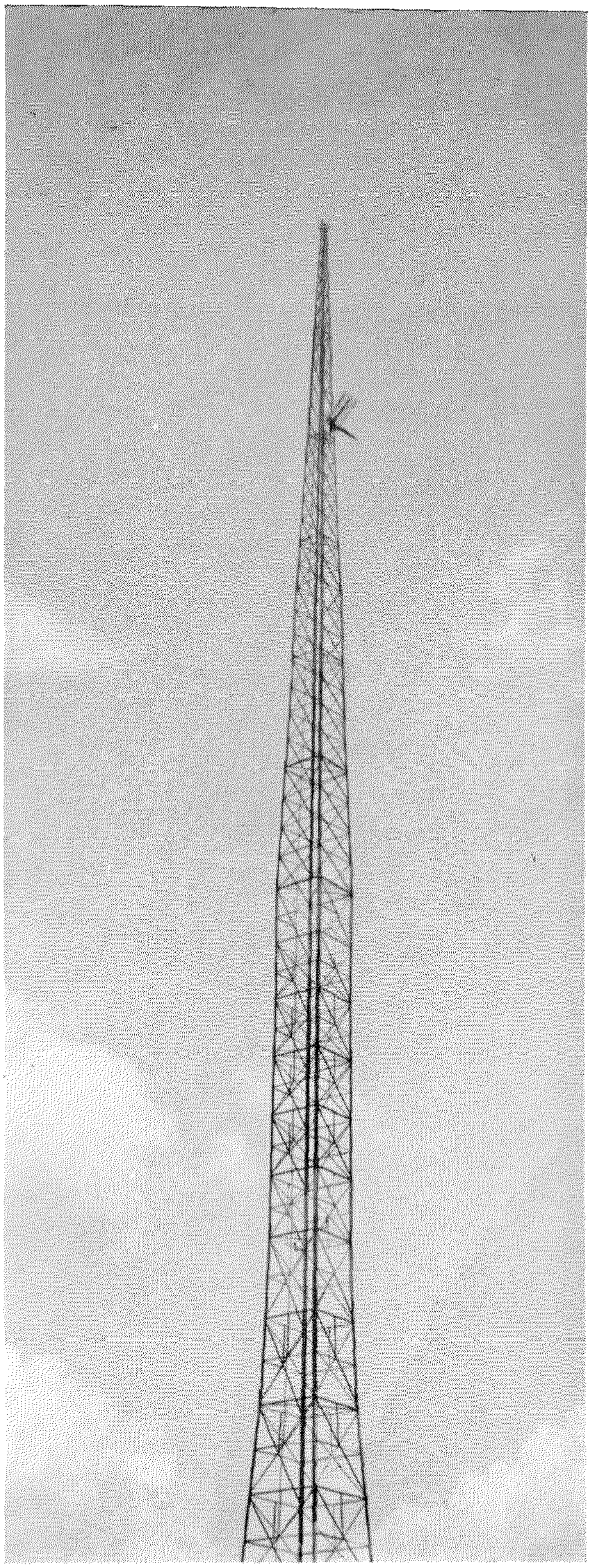

Figure 6—Antennas and portion of supporting tower at Sabana Llana.

The frequency-modulation improvement threshold is 1 microvolt and the noise figure is 7 decibels. The signal normally present at the receiver input terminals is approximately 500 microvolts on the Sabana Llana–Crown Mountain link and approximately 1000 microvolts on the Crown Mountain–Christiansted link.

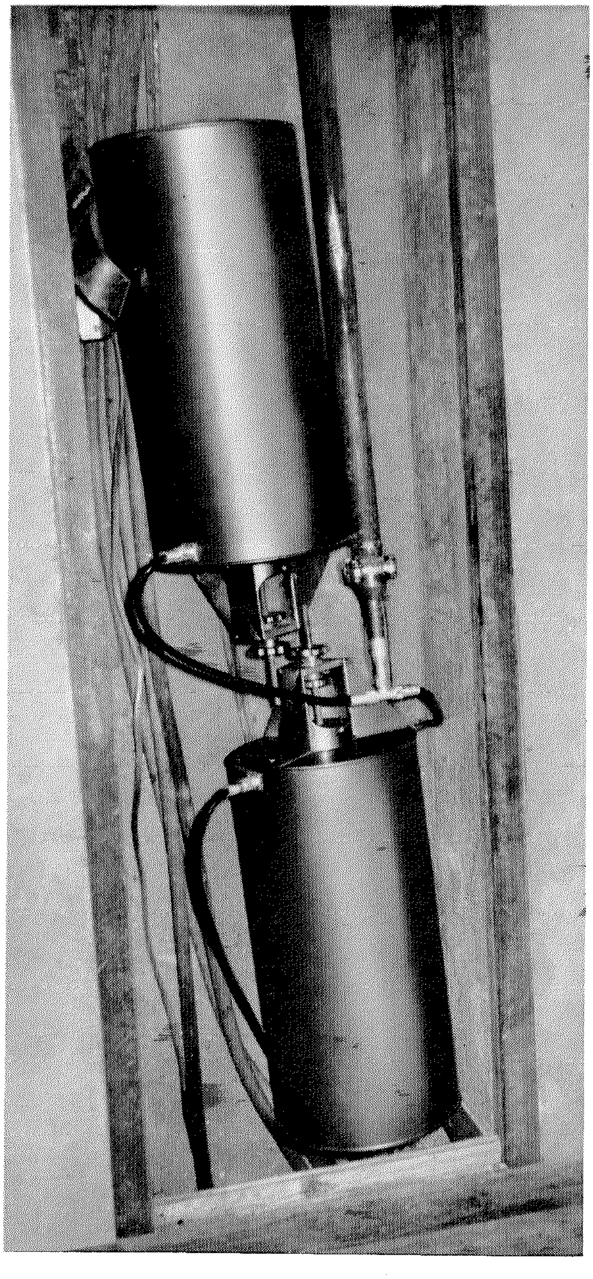

Figure 7—Cavity resonators to keep the transmitted power out of the receiving system at Sabana Llana.

The rack at the left of Figure 3 contains apparatus associated with the cable lines such as repeating coils, volume indictor, filters and jacks. The cabinet to its right contains a remote-control relay panel and three receivers.

The transmitter and receiver together are capable of a system response–frequency characteristic that is flat to within 1 decibel from 200 cycles to 20 kilocycles with less than 1 percent of amplitude distortion and a 50-decibel signal-to-noise ratio per 3-kilocycle channel. Intermodulation between two test tones is 50 decibels down. This performance is considerably better than that of frequency-modulated radio equipment commonly used for mobile services or for single-channel telephone service. The higher standards are necessary to prevent crosstalk on multi-channel links.

Spare radio equipment is provided at each location but no provision is made for automatic switchover in event of failure. The Crown Mountain station is the only unattended location, and for this point a monitoring receiver is provided at the office in Charlotte Amalie which is 4 miles away, with relays for switching transmitters and receivers through the interconnecting cable.

Alternating current supplied by local public utilities is used at all locations, but voltage regulators and gasoline-engine-driven emergency generators with automatic starting and transfer facilities are provided.

5. *Antennas*

Horizontally polarized corner-reflector-type antennas are used. They have a gain of 8 decibels over a half-wave dipole. With the exception of those at Sabana Llana, all antennas are connected to the equipment through $\frac{7}{8}$-inch-(2.2-centimeter-) diameter air-spaced copper-tube coaxial lines, which are filled with dry air at low pressure. Antennas for diversity reception have not been provided because there appears to be no need for them.

Figure 4 is a view of two transmitting and two receiving antennas, mounted on a tower above the station building at Crown Mountain. After this photograph was taken, two of the antennas were moved from the tower to poles about 60 feet (18 meters) away in order to reduce coupling between the Christiansted and Sabana Llana

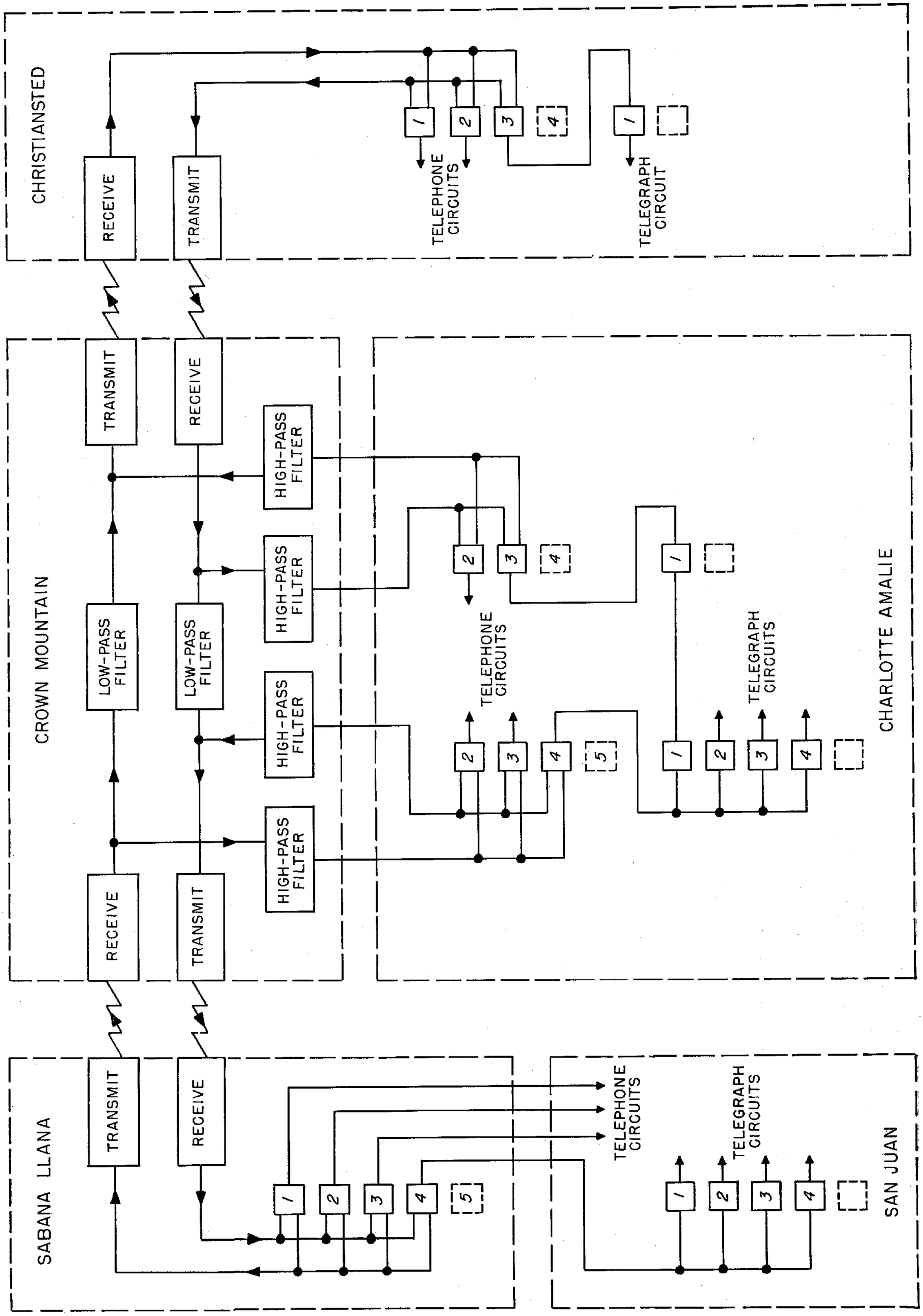

Figure 8—Block diagram of carrier system.

links at Crown Mountain. They may be seen in Figure 5.

Figure 6 shows the single combined transmitting and receiving antenna on a 300-foot (90-meter) tower at Sabana Llana. Since this antenna is 800 feet (240 meters) from the equipment, $1\frac{3}{8}$-inch- (3.5-centimeter-) diameter coaxial line is employed to hold transmission-line losses to 1.6 decibels. At the station end of the line, cavity resonators shown in Figure 7 are used to keep the transmitter radio-frequency energy out of the receiver.

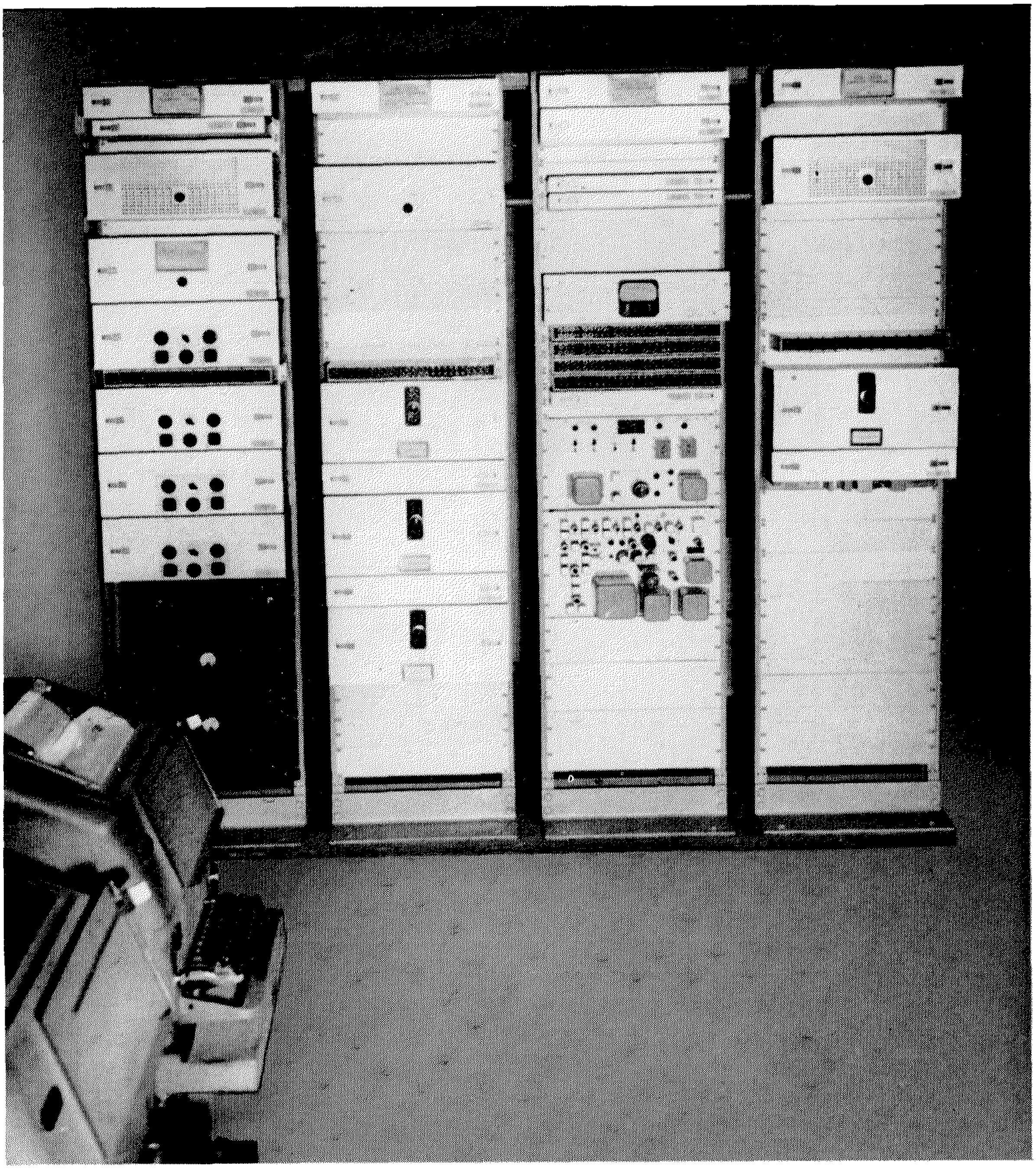

Figure 9—Telegraph carrier equipment at Charlotte Amalie.

At Christiansted, the two antennas are mounted on a 100-foot (30-meter) tower behind the telephone building in the center of the town, where they are exposed to occasional strong ignition radiation from old automobiles on adjacent streets. Interference from this source is generally maintained at 30 decibels or better below speech level by the fairly strong frequency-modulated received signal strength and by the horizontally polarized antennas.

The stations are in the area where hurricanes originate. The self-supporting towers are designed to withstand safely wind velocities of 120 miles (190 kilometers) per hour and they are much heavier than towers commonly used for similar service in the United States. The Crown Mountain tower, which is in the most exposed location, has been strongly guyed in addition.

6. Carrier Equipment

Figure 8 is a block schematic of the system. Channel 1 is on voice frequencies while channels 2, 3, and 4 use low-frequency carriers that are of the same frequency in both directions of transmission. Standard telephone-type single-sideband carrier-suppressed equipment is used.

Channel 1 is transferred at Crown Mountain from one link to the other by means of low- and high-pass filters as indicated in Figure 8 to provide the direct circuit between Saint Croix and Puerto Rico.

Hybrid coils are provided where the 4-wire radio circuits and 2-wire telephone exchanges are interconnected. Two-frequency (1000 cycles interrupted 20 times per second) ringers are provided for signaling between telephone operators.

Standard telegraph carrier techniques are used to subdivide one telephone channel into printer channels. Frequency-shift keying is employed and a channel spacing of 120 cycles is used. Figure 9 is a view of the telegraph carrier equipment at Charlotte Amalie.

7. Operating Results

Over-all unweighted signal-to-noise ratio at the telephone switchboards, which includes noise contributions from lines, carrier equipment, radio apparatus, and crosstalk is about 35 decibels.

During the initial 18 months of operation, service has been available 99 percent of the time between San Juan and Charlotte Amalie, and 97 percent of the time between San Juan and Christiansted. The latter service involves two very-high-frequency links in series.

Principal causes of lost time have been defective vacuum tubes, open fuses due to power surges, faults in land lines, power failures, and noise due to inadequate bonding on antennas and towers.

The corrective measures already taken have improves these initially unsatisfactory conditions and a concurrent reduction in lost time has resulted.

8. Acknowledgment

The successful operation of the system has been due to the work of my associates, K. M. Barbier at Saint Thomas and P. A. Girard and J. Castanera in Puerto Rico.

Microwave Television Radio Relay System*

By O. H. APPELT, KARL CHRIST, AND KURT SCHMID

C. Lorenz, A. G.; Pforzheim, Germany

MICROWAVE relay stations are used to link the television broadcasting transmitters in Köln, Frankfurt, and Neustadt. The main principles that governed the design of these relay equipments are summarized and brief outlines are given of the electrical and mechanical features and of the measurement techniques.

1. *Design Fundamentals*

1.1 Transmission Frequencies

Transmission frequencies, bandwidth, and interchannel spacing were determined in cooperation with the German Post Administration and with the firms manufacturing the equipment. It was decided that the transmission frequencies be in the 1700- to 2300-megacycle-per-second band, which is in accordance with the frequency allocations of the Atlantic City international radio conference.

It appeared necessary to have a bandwidth of 30 megacycles for each channel and allow 60 megacycles for interchannel spacing. This permitted 10 channels or 5 channel pairs to be accommodated in the band. The reasons for the selection of these data have been stated elsewhere[1,2] and will be touched only briefly here.

The basic frequency range was selected primarily with regard to available microwave tubes such as the *2C39*.

The choice of channel width was influenced strongly by the video band being from zero to 6 megacycles and by the use of wideband frequency modulation to obtain the required signal-to-noise ratio for the transmission of sound signals.

For transmission in only one direction, two radio-frequency channels are required to permit the transmitted and received frequencies of a relay station to be offset from each other and thus avoid feedback. If several propagation paths run parallel or cross each other, additional channels are required. Small interchannel spacing enables many channels to be accommodated in a given band.

In multichannel systems, mutual interference may result from a radiating oscillator and insufficient image-frequency suppression in a receiver. Such interference can be eliminated by arranging the oscillator frequency so that the image band falls between the radio-frequency channels. This requirement determines that the minimum spacing of channels be twice the channel width if the oscillator frequency is at one edge of a channel. For this case, the interchannel spacing would be 60 megacycles.

The lowest possible intermediate frequency (5 × 15 = 75 megacycles) was chosen, with consideration for the availability of wideband amplifier tubes. The oscillator frequency for a radio-frequency channel then falls at the far edge of the adjacent channel. In relay stations, there was still the danger of interferences from the transmitter oscillator frequency being radiated into the receiver channel. This danger has been completely removed by an appropriate frequency distribution[3] of the oscillators.

For two channels, the oscillator frequency for the higher-frequency channel is placed 75 megacycles above the channel edge while the oscillator frequency for the lower-frequency channel is placed 75 megacycles below the channel edge. Thus the set-off frequency or the spacing between the two oscillator frequencies is equal to the sum of twice the intermediate frequency and the interchannel spacing (210 megacycles).

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*Originally published under the title of "Die Dezimeterwellen-Richtfunkgeräte der Fernsehübertragungsstrecke Köln-Frankfurt-Neustadt" in *Fernmeldtechnische Zeitschrift*, volume 6, pages 406–410; August, 1953: and *SEG-Nachrichten*, volume 1, number 4, pages 18–22; 1953.

[1] E. Dietrich, K. O. Schmidt, and E. Weingartner, "Neue Fernseh-Rechtfunkverbindungen der Deutschen Bundespost," *Fernmeldtechnische Zeitschrift*, volume 6, pages 220–233; May, 1953.

[2] H. Behling, G. Brühl, and E. Willwacher, "Die Dezimeter—Richtfunkanlage Freda I," *Telefunken-Zeitung*, volume 26, number 98, page 4; 1953.

[3] German Patent 862 320.

1.3 Signal-to-Noise Ratio

The signal-to-noise ratio required[1,4] for the video signal after transmission through 20 relay links was 35 decibels between peak values or 40 decibels between effective values. Taking into consideration the well-known relations among signal-to-noise ratio, transmitter power, receiver sensitivity, antenna gain, attenuation per radio section, and number of relay stations, the following values were laid down as a basis for the design: transmitter power of 5 watts, receiver noise figure of 25 (14 decibels), frequency swing of 12 megacycles peak-to-peak modulation, frequency band of 6 megacycles, and antenna gain of 33 decibels.

1.4 Other Requirements

Other requirements dealt with transmission distortion. They have been described at length elsewhere.[4] The linearity of the characteristics of modulators, demodulators, and video amplifiers should be such that deviations in transconductance do not exceed 10 percent of the mean value over the video range.

Variations of the group-delay time, which distort transients from black to white, should not exceed 0.1 microsecond after transmission through 20 radio link sections. These delay distortions are usually generated in the intermediate- and radio-frequency elements and also in the antenna leads if there are reflections at the antenna and equipment couplings. In these equipments, variations of group delay time in the intermediate- and radio-frequency sections were kept below 0.005 microsecond and the reflection factor at the antenna input terminals was maintained below 2.5 percent.

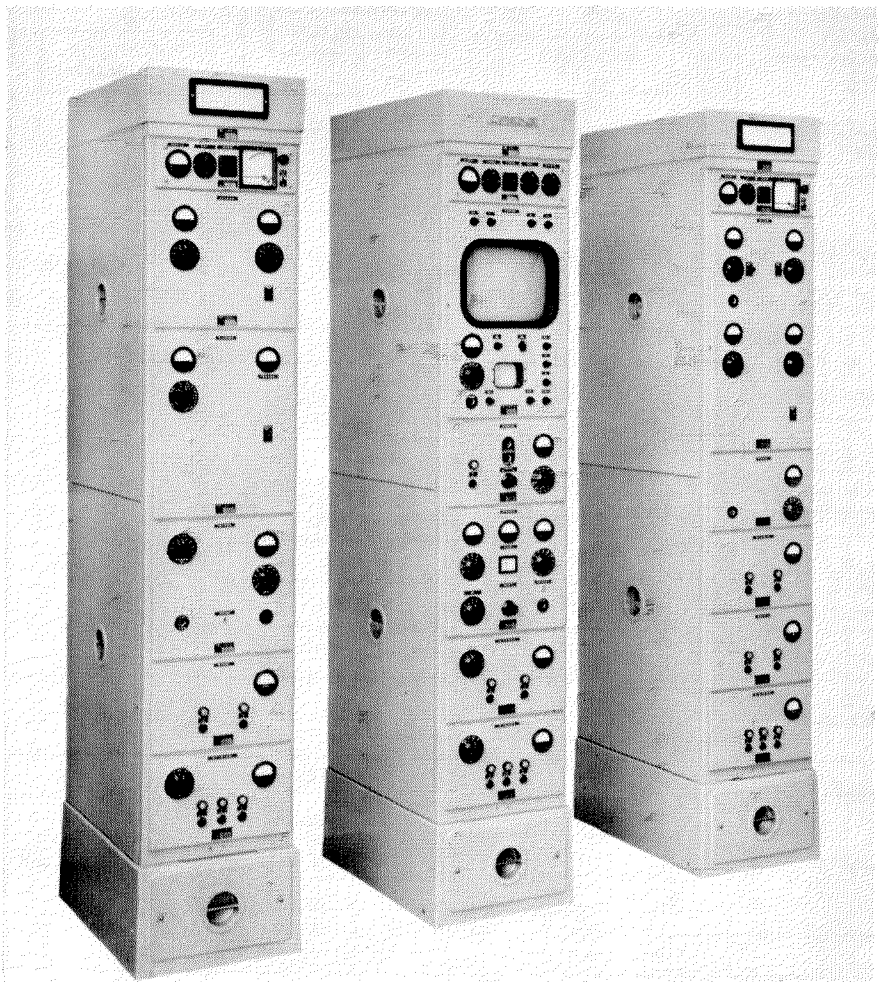

Figure 1—These 3 racks contain the equipment for a terminal station.

2. *Mechanical and Electrical Arrangement*

2.1 Mechanical

Figure 1 shows the terminal-station apparatus mounted in 3 racks. The top panels of the racks contain switchboards with instruments indicating levels and voltages. The antenna switch is mounted under the top cover of the receiver rack. It is remotely controlled and is used to reverse the direction of transmission, which direction is indicated by a lamp signal. The pedestals of the three racks house the blowers.

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[4] E. Dietrich, "Fernsehübertragungsstrecke auf Dezimeterwellen zwischen Hamburg–Hanover–Köln–Frankfurt M." and F. Kirschstein, "Die Übertragungseigenschaften der deutschen Fernsehleitungen," *VDE-Fachberichte*, volume 17, pages 1–12, number V; 1953.

The rack to the right houses the ultra-high-frequency transmitter, the intermediate-frequency power amplifiers, and the necessary power supplies. The left-hand rack holds in its upper part the ultra-high-frequency receiver, beneath which are its intermediate-frequency amplifier, video amplifier, and two power supplies. The center rack accommodates the picture and intermediate-frequency monitor unit together with the modulator and the power supplies.

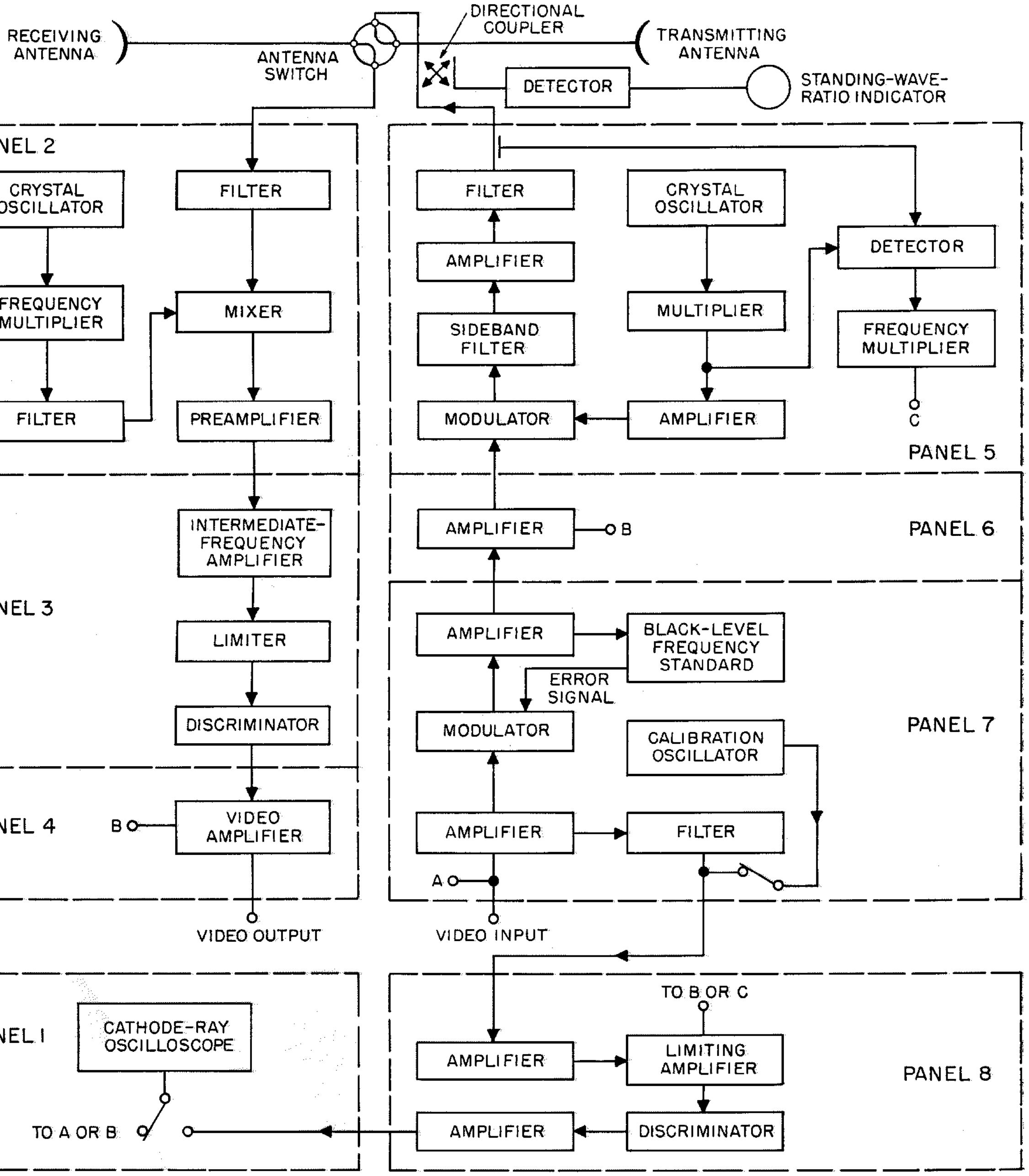

Figure 2—Block diagram of terminal transmitter and receiver.

The relay stations differ from this arrangement only in that they do not have the modulator and its power supply.

The sides of the racks are covered by demountable panels. The individual constructional elements are assembled on a vertical mounting plate that is perpendicular to the front side of the rack. Once the demountable side covers are removed, all parts and the tubes are easily accessible.

2.2 ELECTRICAL

2.2.1 Terminal Station

Figure 2 shows a block diagram of a terminal. In this diagram, the basic elements of the receiver are in vertical alignment in panels 2, 3, and 4 with auxiliary equipment to the left. Similarly, the transmitter elements are aligned vertically in panels 5, 6, and 7, with auxiliary apparatus to the right. This diagram also shows how the various stages are assembled in the individual panels.

The vertical-blanking synchronizing pulse (1.5 volts peak to peak across 75 ohms) is applied to the transmitter through the video-input jack in panel 7. After video-frequency amplification, this signal goes to a modulator that produces an intermediate frequency of 75 ± 6 megacycles. The frequency of 72.36 megacycles corresponds to the black amplitude level. It is compared with a black-level quartz-crystal-generated frequency. In case of a frequency deviation, an error voltage is generated to control the fine tuning of the modulator and bring about a suitable correction.

The modulator panel also includes a calibration oscillator adjustable from 68 to 83 megacycles for alignment of circuits. The synchronizing level is at 69 megacycles, white peaks are at 81 megacycles, and the black level at 72.36 megacycles as mentioned before.

The intermediate-frequency signal obtained from the modulator is amplified in the power amplifier (panel 6) from 0.5 volt to 20 volts and is applied to the modulator of the ultra-high-frequency transmitter (panel 5). There, it is mixed with a quartz-controlled frequency and the desired sideband is filtered out. That sideband is applied to the antenna through an amplifier, an ultra-high-frequency filter, and the antenna switch. The antenna lead is coupled to a measuring instrument for transmission level and a directional coupler to permit the standing-wave ratio to be checked for matching impedances to the antenna. The radiated signal is monitored by a picture monitor.

The received signal passes through the antenna switch and an ultra-high-frequency input filter to the mixer. The oscillator frequency is again generated by a quartz crystal followed by several multiplier stages. The resulting intermediate-frequency signal passes through a low-noise cascade-type preamplifier to the main intermediate-frequency amplifier (panel 3). This amplifier has an automatic gain control that compensates for signal variations as large as 40 decibels. After passing through a limiting stage, the video signal is obtained from the discriminator. This is applied to the video amplifier (panel 4) where it is again amplified, the black-level is adjusted if necessary, and the signal goes to the video output terminals.

The operation of the equipment can be checked by an intermediate-frequency monitor receiver on panel 8 in conjunction with the cathode-ray oscilloscope of panel 1. The former converts the frequency-modulated intermediate-frequency signal into an amplitude-modulated video signal, and the latter shows the television picture on a cathode-ray-tube screen. Below this tube is a switch connecting the cathode-ray tube to a waveform monitor for checking the video level by line and frame sweep frequencies.

2.2.2 Relay Station

A relay station does not differ greatly from a terminal station except for the omission of panel 7 containing the video-to-intermediate-frequency modulator and an arrangement whereby the receiver oscillator frequency is obtained from the transmitter oscillator frequency by a shift of 210 megacycles. The intermediate frequency signal is taken from the intermediate-frequency amplifier and applied to the power amplifier.

2.3 ANTENNAS

The antennas employ paraboloidal reflectors that are 3 meters (9.8 feet) in diameter. One of these is shown in Figure 3. The matching of the

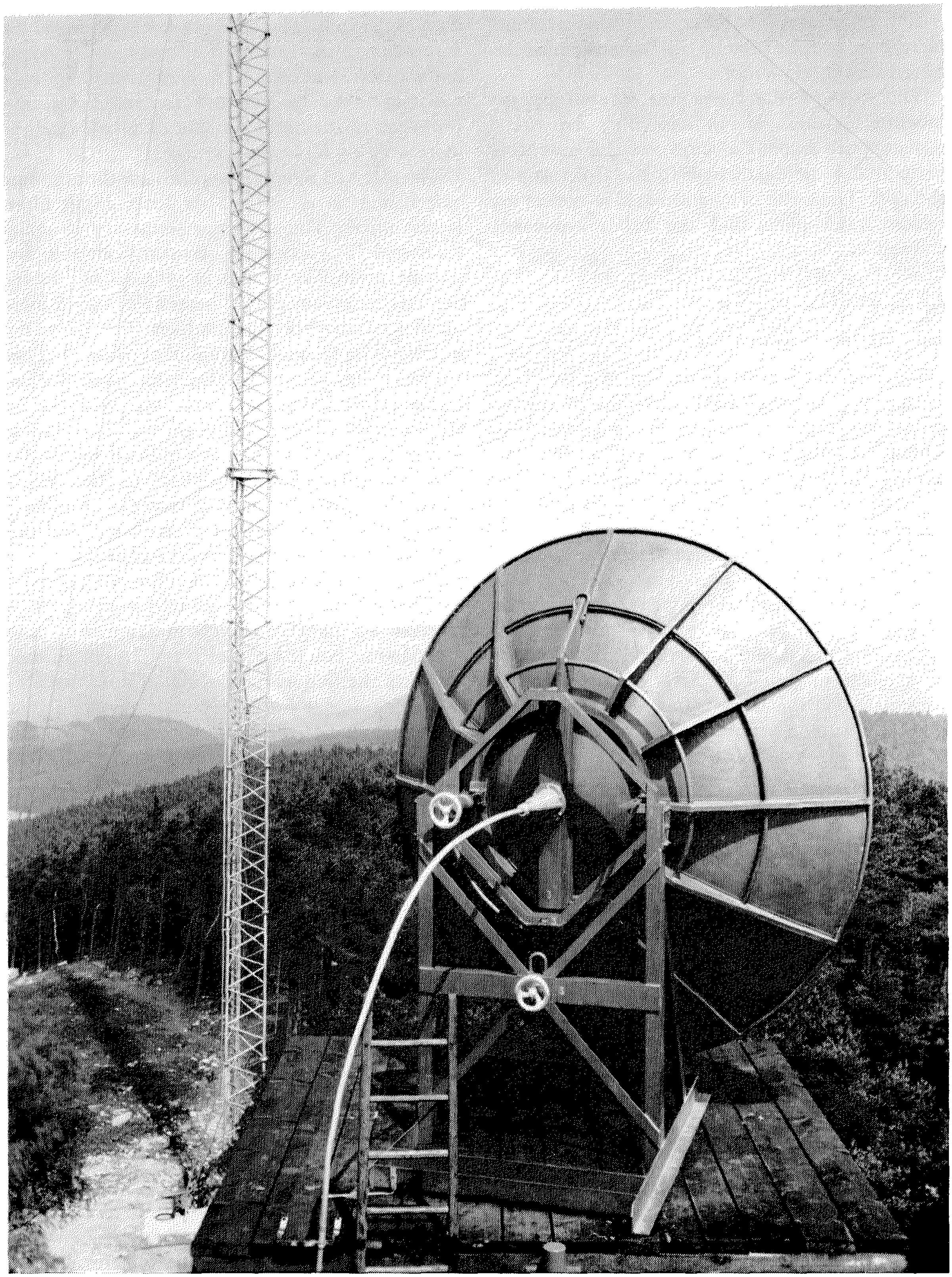

Figure 3—Paraboloidal antennas are used at all of the stations. They are 3 meters (9.8 feet) in diameter.

antenna to the transmission line is better than 0.95. Methods of obtaining this high degree of matching are described later.

3. *Measurements*

3.1 Receiver Sensitivity

The noise source, which is a gas discharge in a coaxial line, has an internal impedance of 60 ohms and a noise temperature corresponding to $70kT_0$ (k = Boltzmann's constant and T_0 = absolute degrees Kelvin); that is, the noise power appearing across a termination of 60 ohms per cycle bandwidth is 18.4 decibels above $1kT_0$. The noise generator is connected to the receiver input terminals via an attenuator having 1-decibel steps starting with zero. During the sensitivity measurement, the limiter in the main intermediate-frequency amplifier is replaced by an equivalent attenuation resistance.

With the gas in the noise source in an un-ionized state, the internal noise power at the intermediate-frequency output of the receiver is indicated by a rectifier and meter that need not be calibrated. The noise source is then energized, an attenuation of 3 decibels is inserted between the preamplifier and the main intermediate-frequency amplifier, and the ultra-high-frequency attenuation is increased from zero until the initial indication at the output rectifier is again obtained. The receiver noise figure is then 18.4 decibels minus the ultra-high-frequency attenuation. (Example: $18.4 - 5 = 13.4$ decibels, which corresponds to a noise power figure of 22.) Measurements taken on factory-produced equipments showed noise figures from 20 to 27.

3.2 Signal-to-Noise Ratio

A well-known method[5] was used to measure the average noise voltage peaks at the receiver output by means of a broad-band oscilloscope and comparison with a square-wave voltage transmitted over the system.

The square-wave generator supplies a voltage producing the full swing of 12 megacycles peak-to-peak. This voltage is applied to the video input of the terminal-station transmitter through a variable attenuator, which is adjusted so that the noise voltage peaks and square-wave pulses are equal in the receiver output. This gives the peak value of the noise, from which the signal-to-noise ratio can be computed. The signal-to-noise ratio depends substantially on the attenuation over the radio link. The measurements coincide well with computed values for signal-to-noise ratio and for a field attenuation over a radio link section of 15.3 nepers or 133 decibels.

3.3 Distortion Measurement

To obtain an idea on the distortion of a signal after having passed through several relay links, the equipments were tested by circulating pulses. The circulating system includes two open-air links, each of 330 meters (1083 feet), between which the channel is changed in a passive relay station. The arrangement is shown in Figure 4.

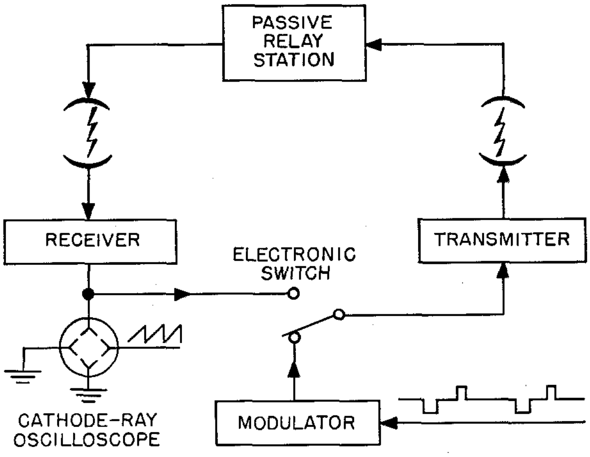

Figure 4—Setup for distortion measurements.

A frame-blanking synchronizing signal is applied to the modulator, which also receives as the contents of a picture a pulse of 1-microsecond duration and a rise time of less than 0.04 microsecond. An electronic two-position switch transfers the pulse to the transmitter intermediate-frequency amplifier and, immediately afterwards, closes the circuit between the intermediate-frequency amplifiers in the transmitter and receiver. The pulse can circulate as many as 20 times during one line sweep.

A gate circuit at the video output of the receiver may be used to inspect the measuring pulse after an adjustable number of circulations.

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[5] C. Cherry, "Pulses and Transients in Communication Circuits," Chapman and Hall, London; 1949.

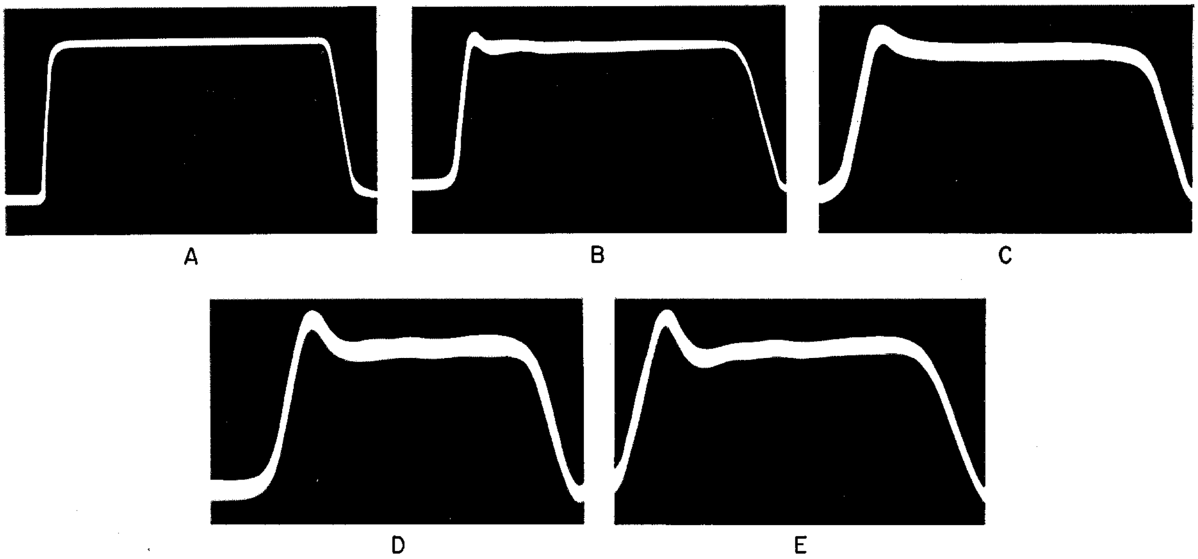

Figure 5—Circulation of square waves through the system of Figure 4. The original 1-microsecond square wave is at *A* and after the specified number of trips around the system at *B* for 1 trip, *C* for 6, *D* for 12, and *E* for 18.

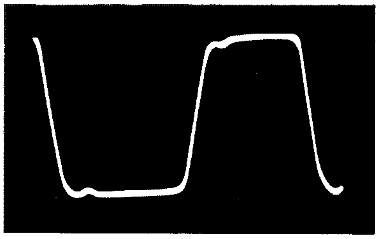

Figure 6—A 500-kilocycle square wave as received at Neustadt from Köln.

Figure 5 shows some examples of such observations: a measuring pulse before being supplied to the circular circuit, and after 1, 6, 12, and 18 runs.[6]

Figure 6 shows as an example of an operational measurement one full square wave at 500 kilocycles transmitted from Köln to Neustadt television station Weinbiet.

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[6] A. C. Beck and D. H. Ring, "Testing Repeaters with Circulated Pulses," *Proceedings of the IRE*, volume 35, pages 1226–1230; November, 1947.

Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles*

By W. P. G. KLEIN

C. Lorenz, A. G.; Esslingen, Germany

TRAVELING-WAVE tubes are commonly used for transmission in microwave relay installations. The output of the receiver and that of a local oscillator are mixed in a silicon diode to produce the modulated radio-frequency wave that is amplified in the traveling-wave tube for retransmission. The maximum output available from a silicon diode mixer is about 0.5 milliwatt. To produce about 5 watts of power for transmission, some 40 decibels or more of amplification must be obtained at the microwave transmitter frequency, requiring two conventional traveling-wave tubes in cascade. This gain can be obtained with a single tube, which will be described. Substantial savings in space, weight, and power are obtained with an increase in operational reliability for the entire relay system.

1. Requirements

In the 4-gigacycle-per-second† range, the power output required in a relay station is about 5 watts. If the tube is to operate near the saturation range, then the saturation gain of the tube must be at least 40 decibels. For reliability under general operating conditions, a gain of about 50 decibels should be obtained in the linear portion of the modulation range. With this high gain, the selection of the operating voltage is of special importance. Employing a normal oxide-coated cathode, the electron gun will produce a converging electron beam at low cathode loading. Therefore, an electron gun with approximately Brillouin flow was indispensable. The operating voltage was fixed at 1400 volts, permitting a satisfactory compromise between tube length and magnetic field. The magnetic field required for focusing the electron beam is about 550 to 600 gausses and can be produced by a permanent magnet.

The gain is significantly smaller at saturation than for small signals in high-gain tubes if no specific measures are applied to correct for this. To keep this difference within tolerable limits, two helixes were employed that have separate direct-current circuits and are coupled only by the electron beam. This results in a further acceleration of the electron beam by the voltage on the second helix, which is somewhat higher than that on the first helix. The difference between small-signal and saturation gains, which was originally 10 to 12 decibels, was thus reduced by approximately 2 or 3 decibels.

A lossy section is placed in the zone between the two helixes. Since the largest amount of power transfer takes place in a small part of the helix near the output end, the second helix should be as short as possible. On the other hand, the axial length of the lossy section should have a certain minimum for low-reflection wideband matching between it and the helix. The efficiency of the tube should not be unnecessarily reduced by the attenuation of the second or output helix; hence the low-loss section of the output helix should contribute a gain of at least 15 decibels. For these reasons, the localized attenuation occurring in the gap between the two helixes is not inserted in the middle of the total length of the two helixes, but towards the output end. The variation of attenuation along the total length is shown in Figure 1.

It is possible to make the attenuation increase approximately exponentially with various exponents. This is recommended to obtain a very-high additional attenuation from as short an axial extension as possible and with low attenuation in the helix. The rate of increase should be easily reproducible as the wave and the beam still interact in the region of moderate attenuation. The longer input helix is responsible for most of the amplification, while the low-loss output helix operates as a final power stage with moderate gain but good efficiency.

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* The contents of this paper were disclosed at a Conference on Electron Tube Research held in Lansing, Michigan, on June 13–15, 1955, under the joint sponsorship of the American Institute of Electrical Engineers and the Institute of Radio Engineers.

† Giga ≡ kilomega = 10^9.

The tube elements are assembled in an evacuated envelope of uniform diameter as may be seen in Figure 2. This imparts good structural rigidity to the tube. The inner diameter of the envelope is determined by the diameter of the electron gun including a ferromagnetic shield inside the tube. As an ordinary oxide cathode is used, the electron gun requires more space than would a simple beam system with a dispenser-type cathode. For this reason, the inner diameter of the tube is 18 millimeters (0.71 inch).

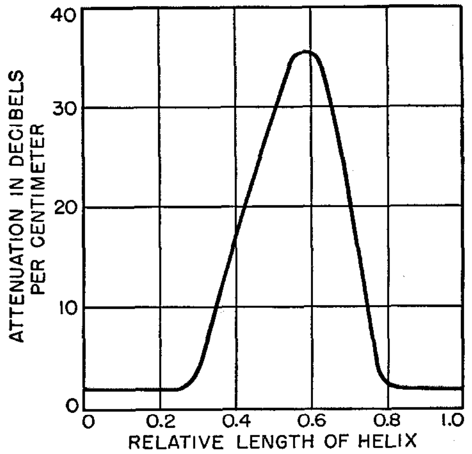

Figure 1—Helix attenuation as a function of relative length. The localized lossy-section attenuation reaches a peak at about 60 percent of the helix structure. The two helixes contribute the horizontal portions of the curve.

Additional precautions must be taken to attenuate fast backward modes but these means must not interfere with the function of the localized attenuation that affects only the fundamental helix wave. A piece of ceramic tubing carries an attenuating layer that has been found experimentally to be of optimum resistivity. The ceramic tubing separates the space between the outer shield and the helix into two heavily attenuated coaxial lines. The diameter of the ceramic tubing is so dimensioned that, for instance, waveguide waves in the space between helix and ceramic tubing do not come into the amplification range of the tube. Due to the high gain at both ends of the localized-attenuation section, the requirements for good

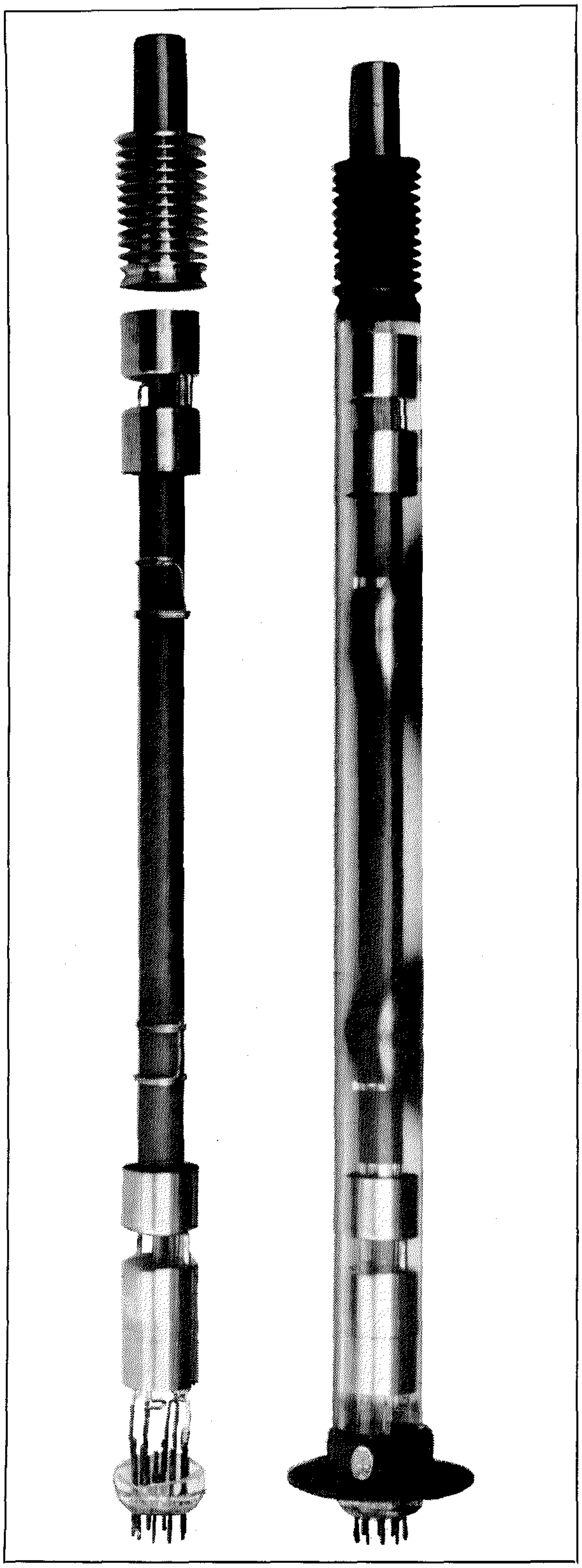

Figure 2—Laboratory sample of *LW53–V* traveling-wave tube. Over-all length is 35 centimeters (13.8 inches).

matching to the helixes are particularly high. This is especially true for the input helix.

To make the input and output impedance independent of variations in the other impedance, the excess of attenuation over gain should not fall below some minimum value, say, 20 decibels, which in this case changes the output impedance by only about 1 percent for full reflection at the input. Therefore, the total reverse attenuation must have a minimum value of 70 decibels. However, the localized attenuation should not require more than a certain portion of the axial length otherwise too large a section of the total helix length of 200 millimeters (7.9 inches) is lost to the process of amplification.

The dimensions of the helix were selected so that the parameter γa for the mean operational frequency is approximately 1.5. Experiments have shown that stability is reduced when a larger γa (for instance, 1.7) was effected by utilization of the same helix diameter and a voltage lower than the rated 1400 volts. With $\gamma a \approx 1.5$, the bandwidth of the tube is about 1 gigacycle between the half-power (3-decibel-down) points of the pass band.

The construction of a localized section of high attenuation and good wideband matching properties without using too much of the axial length of the helix presented a considerable problem. The attenuator is applied to the helix-supporting rods, which are pressed against the helix at several points by elastic clips. The variation of attenuation must be reproducible and capable of being checked by direct-current measurement before assembling the helixes.

The effect of the helix supports on impedance variation should be negligible. Due to the large bandwidth of the tube, an attenuation transition section of about 8 to 10 wavelengths was found experimentally to be desirable although it produces a noticeable difference between small-signal and saturated gain. It is possible to mount two support clips in the region of very high attenuation; the shape of the clips being unimportant. Another pair of support clips—one on the input helix and one on the output helix—are placed in regions of moderate attenuation. Their influence on impedance variations is not so critical in these regions as in regions of low attenuation.

Excepting for a good match to the localized attenuation, the most important requirement is the accuracy of the helix pitch to keep inner reflections to a minimum. In the period of development work, the only helixes available had a noticeable periodic structure. This resulted in heavy inner reflections towards the lower frequencies, that is, below 3.8 gigacycles. The difference in impedance of the tube in the "hot" and "cold" states, that is, with the beam on and off, was large. Only when means were available to reduce the periodic pitch variations of the helix by a special winding technique did this undesirable effect disappear. Also, the electrical stability of the tubes increased. However, even helixes with periodic pitch variations did not show any dips of the gain-versus-frequency curve or appreciable gain variations.

Attention must be paid to the signal-to-noise ratio of a high-gain traveling-wave tube. For a saturation power of 5 watts and a gain of 50 decibels, a noise factor of less than 30 decibels is necessary.

2. *Constructional Details and Measurements*

Figure 3 shows the small-signal gain and the saturation gain versus frequency; Figure 4

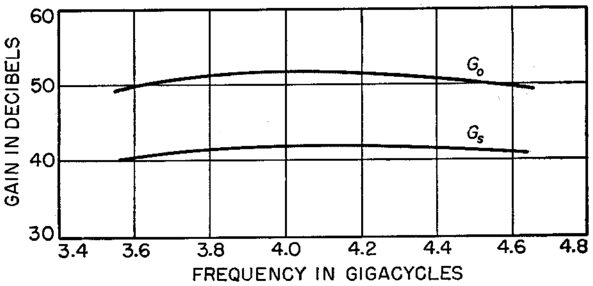

Figure 3—Low-level gain G_0 and saturation gain G_s as a function of frequency (gigacycle ≡ kilomegacycle).

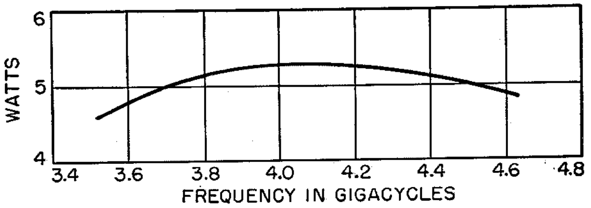

Figure 4—Radio-frequency power output for saturation condition plotted against frequency.

shows the saturation power. The use of an oxide cathode usually requires that the electron gun employ electrostatic focusing. The magnetic focusing of the electron beam within the helix approximates the Brillouin condition.

Measurements revealed that the noise increases with increasing magnetic field in tubes using electron guns having Brillouin flow. The helix current was also higher when the tube was operating near saturation as compared with small-signal performance. If the helix and all other constructional elements of the tube are carefully adjusted, the helix current for negligibly small signals amounts to about 0.3 to 0.6 milliampere for a collector current of 34 milliamperes. At saturation power, however, the helix current rises to about 0.8 to 1.5 milliamperes. The magnetic field required for this current distribution is between 550 and 600 gausses. Figure 5 shows noise and helix current plotted against focusing field strength for two tubes. It will be seen that the noise factor is independent of the focusing field for low values of field. It is only when the field is increased that the noise factor climbs very steeply. The diagram also shows the limits, marked by arrows, of the field strength within which the tube can be operated with maximum output power.

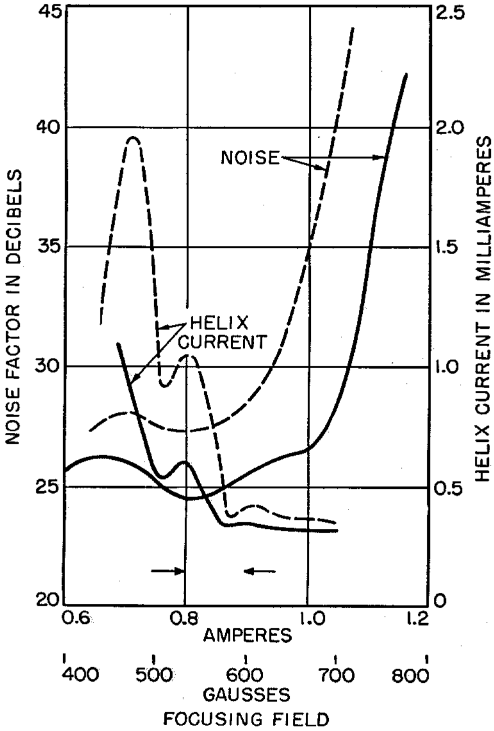

Figure 5—Noise factor and helix current related to focusing field for two tubes, one shown by the solid curves and the other by broken-line curves. The arrows indicate the focusing-field range over which maximum power can be obtained from the tubes.

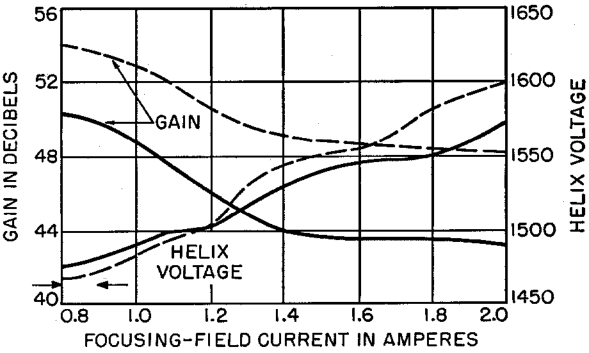

Figure 6—Gain and optimum helix voltage plotted against field current for low-level operation for two tubes. The beam current was 34 milliamperes. The arrows indicate the range of field current over which maximum output may be obtained.

Since the Brillouin condition was necessary for the electron beam, both gain and maximum power output decreased with increasing strength of the focusing field. Figure 6 shows the low-level gain of two experimental tubes versus the magnetic focusing field. The helix voltage was optimized at each measured point and is also shown on the same diagram.

Figure 7 is a plot of the maximum power output of the two experimental tubes against field strength. In the tube corresponding to the broken line, the mounting of the electron gun seems to lack accuracy. This was also indicated in the variation of the helix current with field strength as shown in Figure 5. There are distinct maximum and minimum values of helix current that oscillate about an average value that decreases with increasing magnetic field strength. These peaks are less conspicuous in the curve for the helix current of the other tube. The maximum power output of the first tube plotted against field strength shows the same trend as the helix current, except that a maximum of helix current corresponds to a relative minimum of output power. As to the second tube, increasing field strength results in a steady decrease of the maximum power output.

The effect of the further acceleration of the electron beam by the output helix with its separate direct-current supply can best be observed from the curves of output power plotted against input power of Figure 8. If the direct voltages on both the input and output helixes are adjusted for maximum gain at small signals, curve *A* is obtained. The additional degree of freedom provided by this design permits the output-helix voltage to be adjusted for optimum performance independently of the input helix. The maximum power output for this condition is about 4 watts and the range of modulation is linear up to about 2 watts.

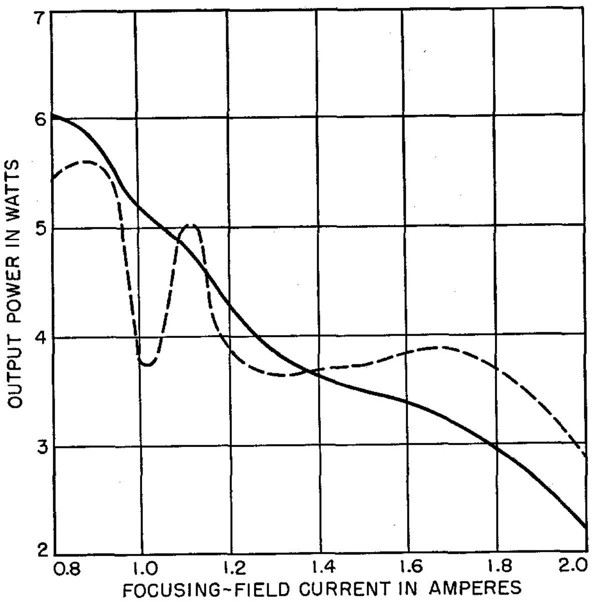

Figure 7—Saturation power output versus focusing field for two tubes.

If both helix voltages are now raised by the same increment to produce maximum power output, curve *B* is obtained. At these higher helix voltages, the gain for small signals is reduced. With increasing input power, the gain of some tubes is also increased and reaches saturation at about 5 to 5.5 watts. If now the voltage of the output helix is raised further above that of the input helix, the maximum power output will be attained and will rise to about 6 or 6.5 watts, as can be seen from curve *C*. In this case, the saturation gain is about 2 decibels higher than for curve *B*. It is seen that curve *C*, after a steady climb, visibly increases its slope, which falls again when the curve reaches the region of saturation. To round off this observation, the gain has also been plotted against power output for these three cases in Figure 9.

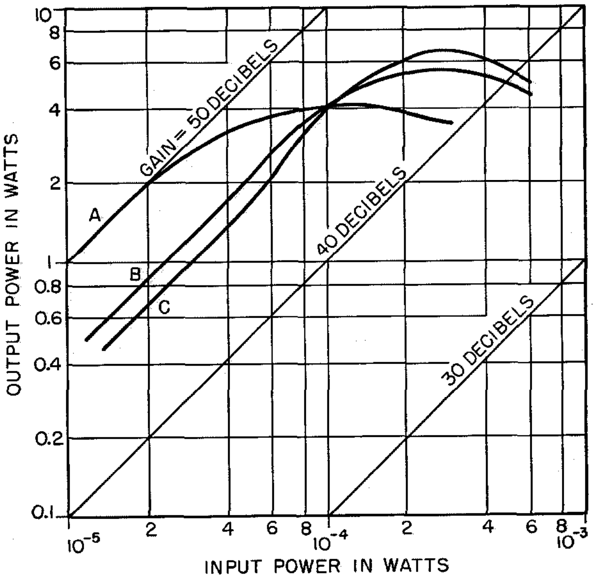

Figure 8—Radio-frequency output power plotted against radio-frequency input power for different values of helix voltages. Curve *A* is for maximum gain under small-signal operation and requires 1470 and 1500 volts on the input and output helixes, respectively. Curve *B* is for maximum power output obtained by simply increasing the helix voltages to 1600 and 1630 volts. By increasing the output-helix voltage to 1670 volts, the further increase in maximum power output shown by curve *C* results. In all cases, the beam current was 34 milliamperes.

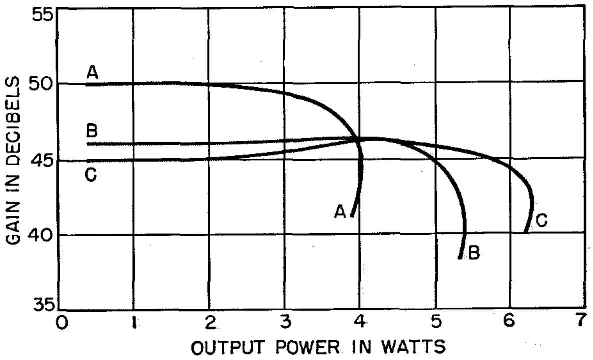

Figure 9—Gain versus output power for the three conditions of helix voltages described in Figure 8.

If an alternating voltage of 50 cycles, for instance, is superimposed on the helix direct voltage, a display may be obtained on a cathode-ray oscilloscope that relates the radio-frequency power output and the direct voltage on the helix. Figure 10 shows such oscillograms; curve *A* is for low-level operation and curve *B* is for saturation operation. These two curves correspond to the conditions outlined for the similarly identified curves of Figure 8. Curves *C* and *D* were obtained when the output-helix voltage was increased further to obtain maximum power output.

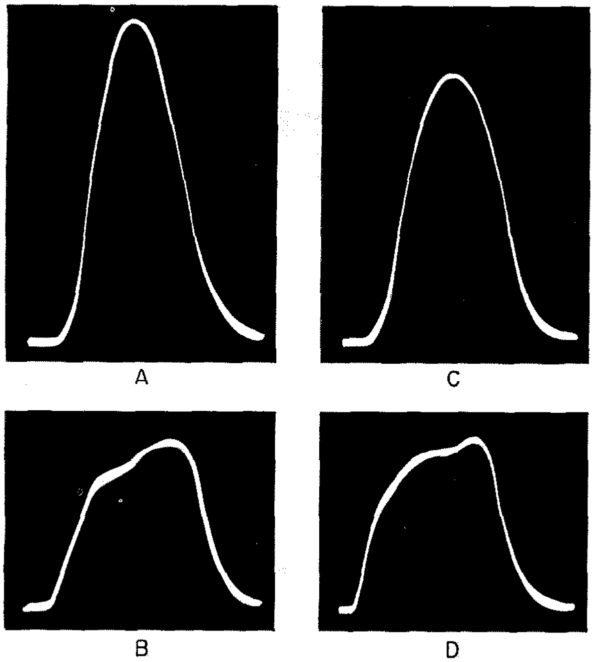

Figure 10—Oscillograms of radio-frequency output power as a function of the helix voltages. An alternating voltage is superimposed on the direct voltage applied to the helixes. *A* and *B* correspond to the conditions for those curves in Figure 8. Curves *C* and *D* were obtained by increasing the output-helix voltage to obtain maximum power output.

In high-gain traveling-wave tubes, it is essential to know about internal reflections in addition to gain, bandwidth, saturation power, noise factor, and stability. There are possibilities for reflections not only at the points of transformation from the helix ends to the waveguides, but also along the helix proper, at the support clips, and at the two transitions to the localized attenuating section. A sensitive method of measuring these reflections is the measuring of the difference in input impedance and output impedance in the "hot" and "cold" states of the tube.

To determine the internal reflections, the waveguide should be sufficiently well matched to the input helix so that there are no reflections at this junction. The impedance can now be observed while the helix voltage is varied. A clear picture of the reflections in the interior of the tube is obtained if a suitable alternating voltage of power frequency is superimposed on the helix direct voltage. An oscilloscope will show a horizontal line if there are no reflections or if the tube is not in operation. The length of this line is proportional to the alternating voltage superimposed on the direct voltage of the helix. If reflections occur at the output end of the helix or elsewhere in the tube, then the horizontal line will be deformed. The ordinate corresponding to each value of the helix voltage will represent a certain degree of reflection.

Two typical examples of such deformations, which can assume sinusoidal shapes, are shown in Figure 11. The amplitude peaks of these oscillations can offer a clue to the magnitude of the internal reflections and the periodicity to the point at which the reflection takes place. Measurements of this type are made at the input and output of the tube and for as many frequencies as possible. At the opposite end of the tube, a waveguide branch can be installed to obtain data on the amount by which the attenuation exceeds the gain.

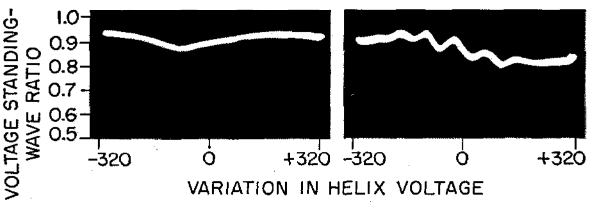

Figure 11—Oscillograms of variations in output impedance caused by internal reflections in two tubes. The abscissa values are for the instantaneous amplitudes of the alternating voltage superimposed on the normal direct voltages on the helixes.

If the maximum impedance deviations caused by reflections in the tube interior are plotted as degrees of matching against frequency, the curves shown in Figure 12 are obtained. These curves are for three different tubes.

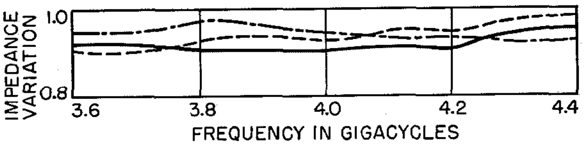

Figure 12—Variation of impedance caused by internal reflections plotted against frequency for three tubes. The *impedance variation is the minimum ratio of the impedance* with the beam on to the impedance with the beam off.

The input and output impedances of the tube should now deviate from the characteristic impedance of the waveguide coupled to it by only a very small amount throughout a fairly large frequency range. Thus, the tube can be very well matched within a limited frequency range of, say, 30 megacycles to the coupling circuits by means of additional selective tuners. It is well known that this is particularly true for the output impedance. The output impedance of a tube, plotted on a Smith chart, is shown in Figure 13. With the tube switched on and without additional tuning means, the voltage standing-wave ratio is not greater than 1.4 between 3.8 and 4.4 gigacycles.

Due to the large level differences between input and output circuits, the whole amplifier or repeater unit should be carefully shielded to avoid stray couplings that might cause additional undesired impedance variations, and, hence, group-delay time distortions.

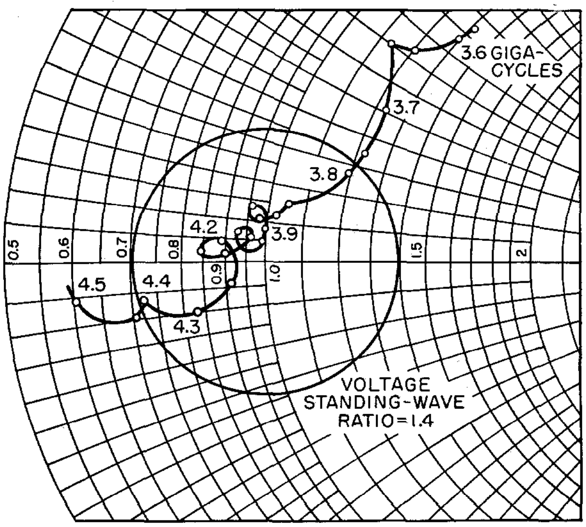

Figure 13—Output impedance as a function of frequency. The frequency values indicated are in gigacycles.

From a recent paper[1], it was learned that the noise behavior of traveling-wave tubes with electron guns producing Brillouin flow in general is not as good as for a tube with a partially shielded gun. Recent measurements show that the noise increases with increasing radio-frequency input power near saturation in tubes with Brillouin flow. Figure 14 shows a plot of the noise figure against radio-frequency input power for three tubes using shielded guns. The noise increases up to some 40 decibels, which is intolerable in the case of a high-gain tube.

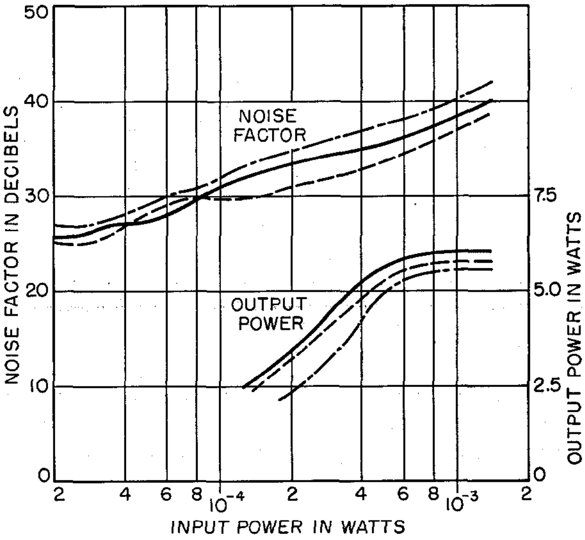

Figure 14—Noise factor versus radio-frequency input for three tubes. The beam currents were 34 milliamperes.

In view of this performance, a modified partially shielded electron gun was developed. The noise factor of three tubes with this gun is plotted against magnetic focusing field strength in Figure 15. It can easily be seen that in this case the noise is almost independent of the field. The measured noise factor plotted against increasing radio-frequency input power is shown in Figure 16. The noise factor of these tubes at low level is approximately 3 to 4 decibels lower than that of tubes with Brillouin flow. Furthermore the noise factor only increases about 4 to 5 decibels at saturation and its value in general remains below 30 decibels. This occurs if the helix voltage is approximately adjusted for maximum gain for low-level operation.

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[1] J. T. Mendel, "Magnetic Focusing of Electron Beams," *Proceedings of the IRE*, volume 43, pages 327–331; March, 1955.

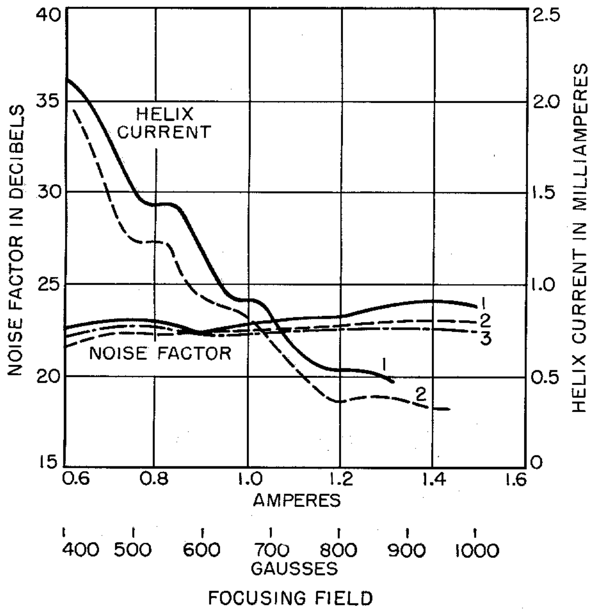

Figure 15—Noise factor and helix current plotted against focusing field for three tubes employing partially shielded electron guns. The beam currents were 34 milliamperes.

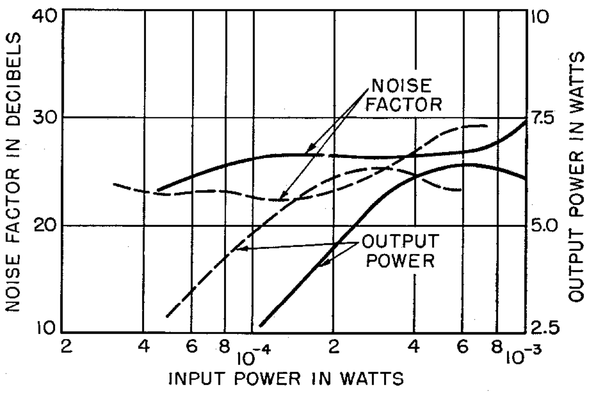

Figure 16—Noise factor and radio-frequency output power plotted against radio-frequency input power for two tubes having partially shielded electron guns. The beam currents were 30 milliamperes.

The tubes with partially shielded electron guns were found to be less critical in operation than tubes with Brillouin flow. Finally, it was observed that the saturation power output of a traveling-wave tube with a partially shielded gun was higher than that of a tube with a Brillouin-type gun with the same electrical data. The average value was 7.5 to 8 watts instead of 6 to 6.5 watts for the tube with Brillouin flow. The gain was also a few decibels higher.

Sampling Procedures on Finished Chassis and Equipment*

By HENRY KNAPP and CHARLES HARTMANN

Federal Telephone and Radio Company, a division of International Telephone and Telegraph Corporation; Clifton, New Jersey

UNRESTRAINED demand for high quality and unquestioned reliability is something that everyone in the electronic industry is experiencing today. It cannot be said that these demands are unjust or impossible for all too often the performance of electronic equipment becomes a matter of life and death.

How, then, does a company build equipment possessing these quality characteristics and how do manufacturers assure their customers and themselves that performance of the equipments in the field will be satisfactory before the fact?

Certainly first-rate production techniques and processes must be installed and maintained. But the watchdog of that maintenance is usually inspection of some kind.

The employment of sampling as a technique to replace 100-percent inspection provides additional inspection coverage for the same inspection effort. We accept this as a statement of fact in applying the sampling technique to inspection of completed equipment. Obviously, however, the use of sampling procedures on completed chassis does not in itself warrant blind faith in the product.

The inspection and control of purchased components must be reliable, and statistical-quality-control techniques should be used for incoming inspection to the fullest extent. Additional information beyond that obtained from routine test and inspection pertaining to performance, reliability, and life on critical items is the responsibility of special laboratory groups within the quality-control organization.

The need for good quality control during assembly and soldering cannot be overlooked. Of course, methods and procedures will vary from job to job as well as from company to company, but probably the greatest need in the assembly area is for immediate information about any trouble. It is wise to use wherever possible a system that pin-points the location of the trouble, the type of fault, and the responsible operator or operation. We call it the point-to-point or operator-identification technique. Figure 1 is a point-to-point tally sheet or chart diagram in which each component and soldered joint is identified and listed in sequence. The inspector notes a defect in the proper spot on the tally sheet and

Inspector:

Date:

RT 70 IF CHASSIS POINT-TO-POINT WIRING INSPECTION

	C-156				Ground				O-101												C-157			Ground					Std.-off					TOTALS
	A □	B ⌓	C △	G —	G8	G9	G10	G11	1	2	3	4	5	6	7	8	9	10	11	12	A □	B ⌓	G —	G12	G13	G14		G15	TP5	TP6	TP7	TP8		
																										L	C							
No Solder																																		
Cold, Rosin Solder																																		
Insuf. Excess Solder																																		

Figure 1—Sample section of a wiring-inspection record. Each soldering point is identified on a wiring diagram. The symbols for *C-156* and *C-157* will be recognized as those used for the terminals of electrolytic capacitors. Several of these report sections will be needed for most equipments.

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* Reprinted from *Proceedings National Symposium on Quality Control and Reliability in Electronics*, pages 15–18; 1954. Presented at the Symposium in New York, New York, on November 12, 1954.

the foreman can tell at a glance the source of the trouble and take corrective action.

Such steps as these: incoming inspection control, specialized testing and research-laboratory examinations of purchased components, and pinpointing of area and operator defects on the production line are necessary prerequisites to successful sampling procedures in the inspection of completed chassis.

1. Sampling

Sampling as a technique to replace 100-percent inspection is not a new idea. It has been used in many places and for many different products. In a great many of these instances, the product was fairly simple and inspection error could be kept at a minimum. Sampling of completed chassis and subassemblies has been carried out previously by various government agencies, but in almost all cases this sampling inspection was preceded by a normal 100-percent inspection. The extensive sampling of completed assemblies direct from production is a comparatively new approach especially in complex electronic equipment that by its inherent design and its requirements has a high potential for error.

The chief objective of this sampling procedure being to insure a consistently high quality of manufacture that produces finished equipment of assured reliability, repeated emphasis must be placed on the self-evident observation that *quality must be built into the equipment, it cannot be inspected in.*

An integral part of a production process is the operator on the line. The chief requirement for high-quality work is the desire on the part of the operator to produce that quality. The operator must want to produce good work or no plan ever designed will improve and insure this quality. Together with this desire, the operator must be well trained, must have proper tools and working conditions, and above all must know the difference between good and poor workmanship. Too often, inspection procedures are such that inferior workmanship is never brought to the attention of the operator responsible for it.

An obvious and very effective method of approaching this goal is to return defective work to the operator responsible for it. The faults of a given operator become evident from the defects listed on the inspection tally sheet and from the various repair operations. Moreover, if the defect is a direct result of some previous operation or a component, the operator can inform the foreman immediately and corrective action can be taken. If the operator is working on an incentive plan of the bonus type, the bonus earned will be closely related to the effectiveness of the work done since high-quality work will not be returned for time-consuming repair.

It may be noted that any manufacturing organization is in business for one reason, profit. A high-quality product at high cost will not enable the company to remain in competition very long. Thus the other objective of this sampling procedure is to lower the production cost of the equipment. This objective can be accomplished in part by lowering the amount of scrap and rework and by lowering the over-all inspection cost.

One of the primary considerations in installing a sampling plan is the type of production process to be sampled. One common process to be utilized is the ordinary straight-line procedure where an inspection station is placed at the end of one or more production operations and the inspector is responsible for examining the work from these specific operations. At this point, the responsibility for any defective work must be clearly established or the entire effectiveness of the plan will be lost. It is also possible for the work of each individual operator to pass but for a lot to be rejected because of an accumulation of minor defects. The problem then becomes one of training operators to refine the workmanship techniques.

A second type of process that has proved adaptable for sampling is the batch type. A group of identical chassis are mounted on a multiple-position rotating jig, commonly referred to as a "lazy Susan." A single operator is responsible for performing all of the assembly and soldering operations necessary for completion of each chassis. At the completion of the last operation on the last chassis on the wheel, the entire batch is submitted to the inspector. This type of process has several advantages.

A. All defects can be attributed to a single operator.

B. The lot size is constant and no opportunity for mixing lots is possible.

C. Minimum handling occurs before inspection.

D. The operator can be informed immediately of defects.

A modification of this lazy Susan process is the use of two wheels. In this case, the operator continues production on the second wheel while the work on the first is being inspected. At the completion of the second wheel, the operator makes necessary repairs to the first batch. Until all repairs are reinspected and cleared, no new units may be started on either wheel. This condition reduces the bonus earning time of the operator, who will avoid it.

Difficulties may be encountered in the installation of the sampling procedure, chiefly centered in the attitude of affected personnel. The operator will realize quickly that she no longer has the protection given by 100-percent inspection; she can no longer continue to produce units at the maximum rate with no personal concern over quality. Penalty for careless work given in the form of rework and repair time taken out of production, puts the emphasis for the operator in the place where it hurts, in her pay check. The inspector, on the other hand, may see extensive sampling as a means of greatly reducing inspection personnel by the substitution of "technical services." The production foreman may see in the plan another harebrained idea thought up by the statistical engineer, which will prevent him from meeting his monthly production schedule. Last but not least, the inspection foreman may complain that sampling means a lot more paper work in maintaining effective data and charts and this will overload his already-crowded schedule. To each of these groups must be demonstrated with clarity and emphasis the fact that work originally produced at a high level of quality is the answer to their doubts and fears.

The matter of returning defective work directly to the operator responsible rather than to an especially designated repair operator, as is the custom in many plants, may require some formal negotiation. While some local compromise arrangements may be in order so long as the principle of operator responsibility is not compromised, in the long run successful selling of the idea supplies the solution to all phases of the problem.

Well-planned administration and control of sampling procedures can prevent breakdown of the system because of occasional human weakness. An inefficient inspector can accept a bad lot and claim that defects found in performing subsequent operations were all located in the uninspected portion of the lot. An ill-advised foreman can manage in an extreme case to hold a rejected lot aside and resubmit it at a later date without rework, hoping that selection of different units for a sample will enable the lot to pass. However, controls such as serial numbers on units, the listing of all serial numbers in a given lot on the inspection record, stamping of the items selected for the sample, and signatures on submission tickets can regulate the system successfully.

2. Classification of Defects

The most-important consideration in any sampling is the classification of defects. Unless all inspectors classify each particular type of defect in exactly the same way, no acceptable quality level or process average can be realistic. A classification is the enumeration of possible defects of the unit of product according to their importance. A defect is any deviation from requirements of the specification, drawings, or purchase-order descriptions, and any changes pertaining to any of them made in the contract or order. Defects are customarily classified in the following manner.

2.1 Critical Defects

A critical defect is one that judgment and experience indicates could result in hazardous or unsafe conditions for individuals using or maintaining the product. For major end-item units, a defect that could prevent the performance of the tactical function of the end item is classified as critical.

2.2 Major Defects

A major defect is one, other than critical, that could result in failure or materially reduce the usability of the unit of product for the intended purpose.

2.3 Minor Defects

A minor defect is one that does not materially reduce the usability of the unit of product for its intended purpose, or is a departure from established standards but having no significant bearing on the effective use or operation of the unit.

2.4 CONTROL DEFECTS

A control defect is any departure from good workmanship that will in no way affect the end use of the product.

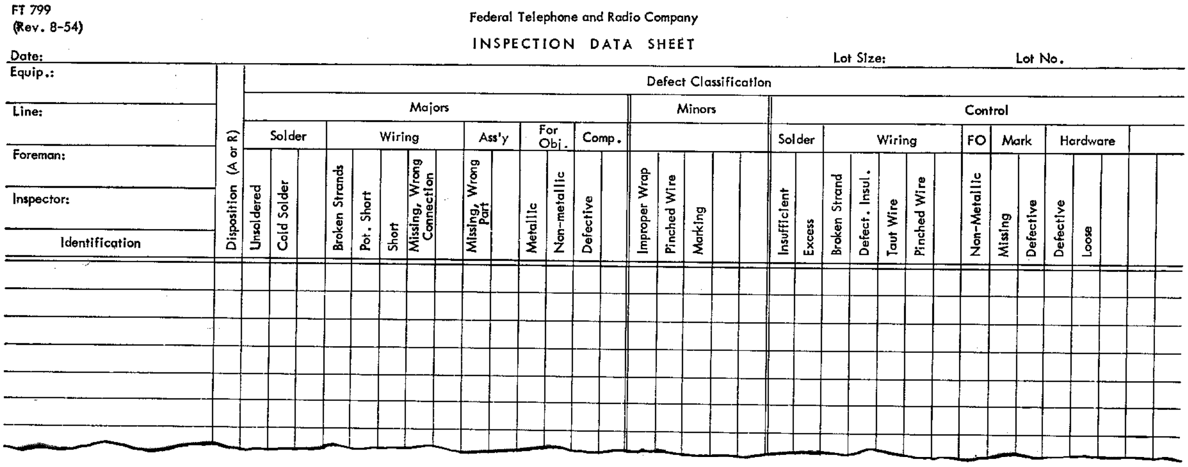
It may be noted from Figure 2 that there is a great similarity between the above classification of defects and that used by various government agencies. Since many of the units produced in our company are sold to the government, it is natural that the defect classification should be in accord, otherwise a lot that was acceptable by the company might be rejected by the government.

The classification will list all categories of defects relevant to the mechanical assembly of the equipments under consideration. Within each category, as many specific defects as possible should be described and classified or weighted in terms of degree of effect on service, maintenance, and performance of the equipment. For example, excess solder on a joint would be named in the category of soldering, and would be defined as a major defect where the amount of solder was sufficient to cause a possible short-circuit. It would be called a minor or control defect, however, when the amount of surplus solder involved only a question of appearance or workmanship standards. The more complete the listing of

TABLE 1

CLASSIFICATION OF DEFECTS

Category	Major	Minor	Control
1. Soldering			
1.1 Unsoldered joint	X		
1.2 Insufficient solder	X		X
1.3 Excess solder			X
1.4 Cold solder	X		X
2. Wiring and Cabling			
2.1 Broken strands	X		X
2.2 Poor lead dress	X	X	X
2.3 Defective insulation	X	X	X
2.4 Taut wire	X		X
2.5 Improper wrap	X	X	X
2.6 Improper crimp			X
2.7 Potential short-circuit	X		X
2.8 Short-circuit	X		
2.9 Insulation in solder			X
2.10 Missing or wrong connection	X		
3. Assembly			
3.1 Fastening and hardware	X	X	X
3.2 Improper mechanical operation	X	X	X
3.3 Improper assembly	X	X	X
3.4 Missing, wrong part or component	X	X	X
3.5 Lubrication	X	X	X
3.6 Pinched wire	X	X	X
4. Foreign Objects			
4.1 Metallic	X	X	
4.2 Nonmetallic	X		X
5. Finish			
5.1 Improper finish	X	X	X
5.2 Defective finish	X	X	X
6. Marking and Identification			
6.1 Safety, warning, or operational	X		X
6.2 Identification	X	X	X
7. Components			
7.1 Defective	X		X



FT 799
(Rev. 8-54)

Federal Telephone and Radio Company

INSPECTION DATA SHEET

Date:
Lot Size:
Lot No.
Equip.:
Line:
Foreman:
Inspector:
Identification

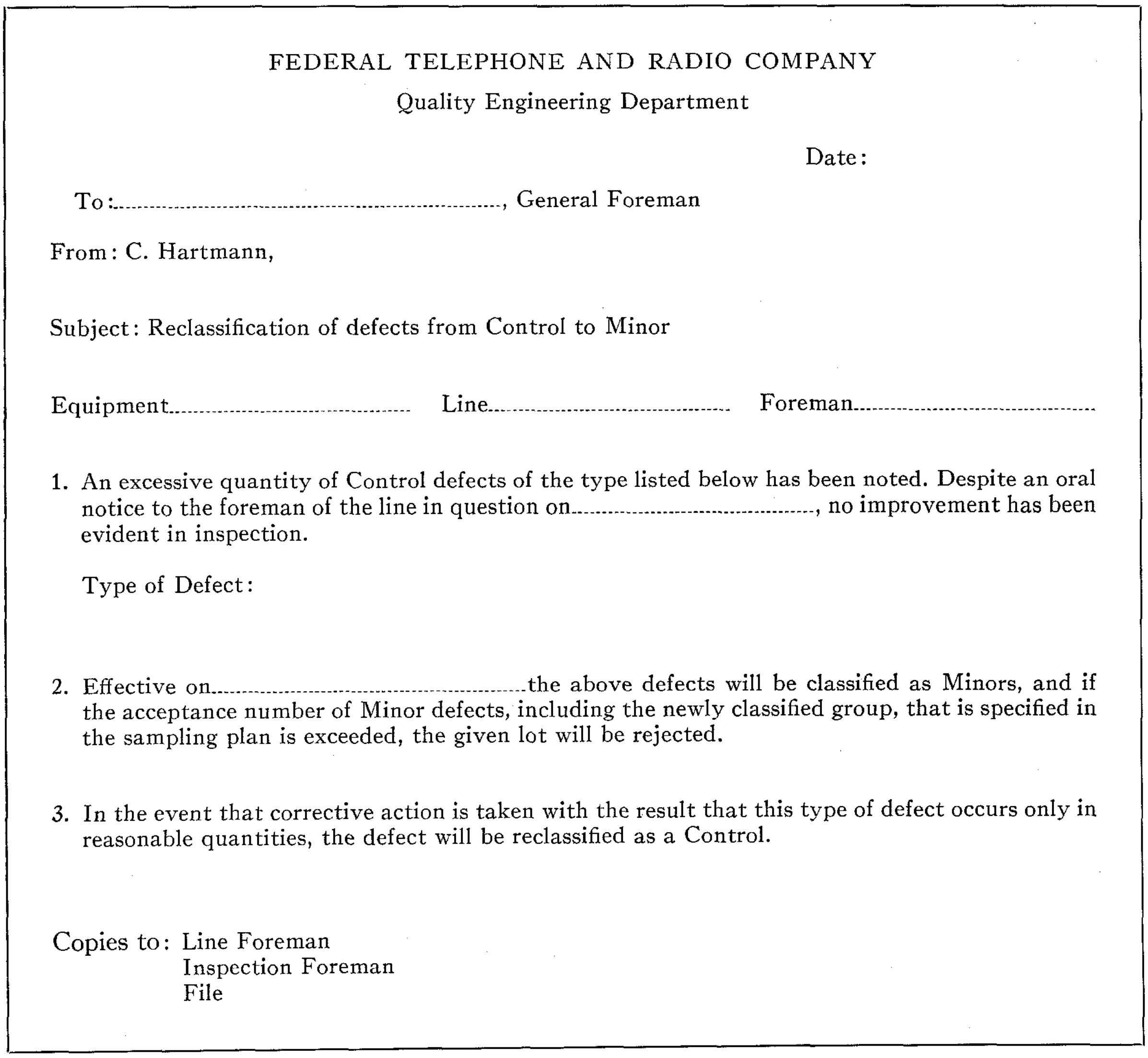
	Defect Classification																																	
	Majors														Minors					Control														
	Solder			Wiring					Ass'y		For Obj.		Comp.							Solder		Wiring					FO	Mark		Hardware				
Disposition (A or R)	Unsoldered	Cold Solder		Broken Strands	Pot. Short	Short	Missing, Wrong Connection		Missing, Wrong Part		Metallic	Non-metallic	Defective		Improper Wrap	Pinched Wire	Marking			Insufficient	Excess	Broken Strand	Defect. Insul.	Taut Wire	Pinched Wire		Non-Metallic	Missing	Defective	Defective	Loose			

Figure 2.

defects, the more complete and accurate will be the results of inspection. Table 1 is a classification of defects.

One of the most-important phases of the program is the training of inspectors. In our instance, selected quality-control personnel were utilized as instructors. Since most of the inspectors involved were experienced, the training was devoted to an understanding of the function and methods of sampling, the classification of defects, and to the meaning of acceptable quality levels and like terms. One difficulty encountered in the training program was the concept that control defects must be recorded but would not be repaired. The inspectors reasoned that if the defect was not of sufficient importance to need repair, it was not important enough to take time to record. Only when it was emphasized that a record of an increasing trend in control defects pointed to a decreasing quality level and that repair work in a complicated chassis could create new defects like burned or broken wires, did the inspection group accept the validity of the concept.

The recording of large numbers of control defects notifies production personnel that steps must be taken to improve quality. If no improvement is noted, control defects should then be



FEDERAL TELEPHONE AND RADIO COMPANY
Quality Engineering Department

Date:

To:.., General Foreman

From: C. Hartmann,

Subject: Reclassification of defects from Control to Minor

Equipment.................................... Line.................................... Foreman....................................

1. An excessive quantity of Control defects of the type listed below has been noted. Despite an oral notice to the foreman of the line in question on...................................., no improvement has been evident in inspection.

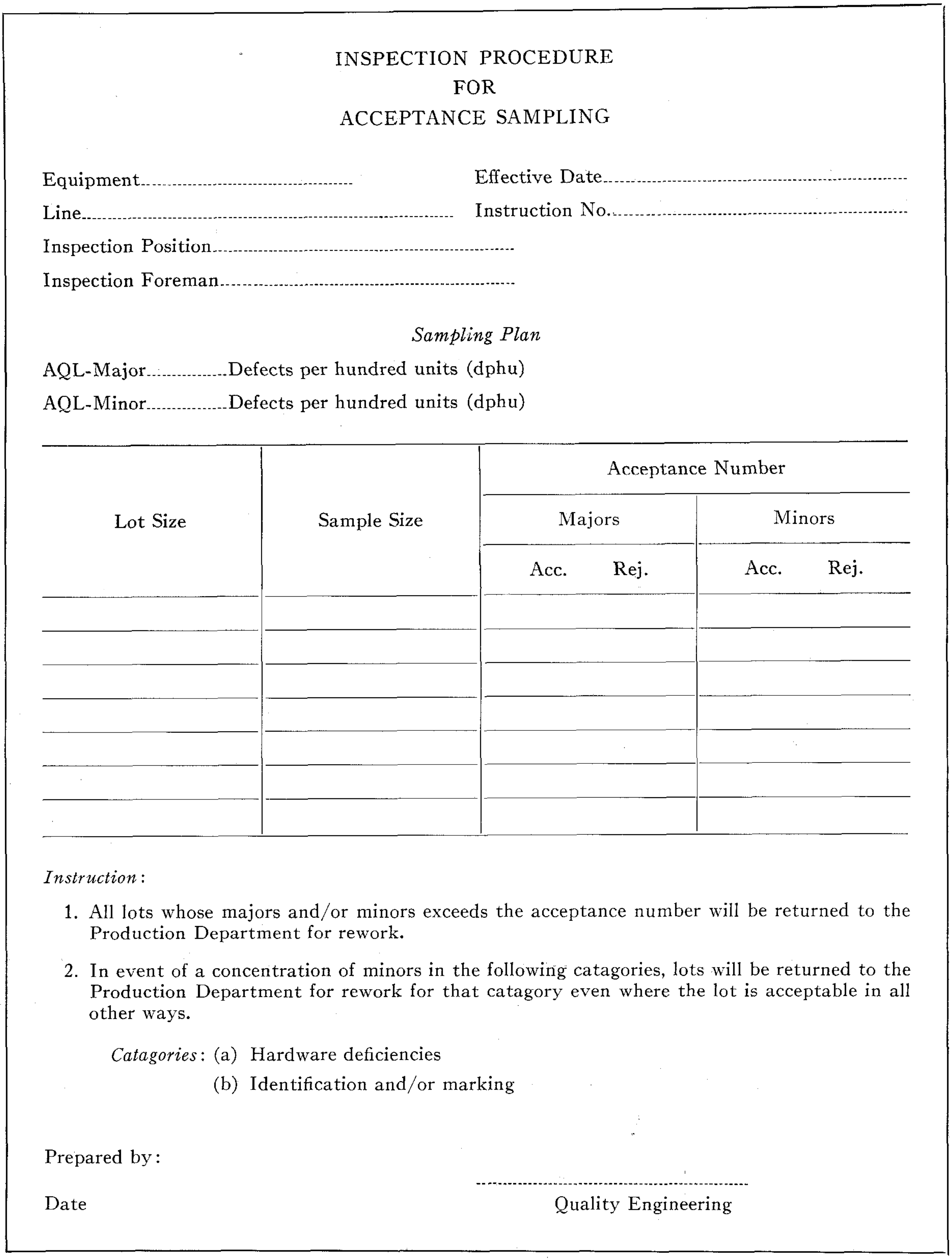
   Type of Defect:

2. Effective on...the above defects will be classified as Minors, and if the acceptance number of Minor defects, including the newly classified group, that is specified in the sampling plan is exceeded, the given lot will be rejected.

3. In the event that corrective action is taken with the result that this type of defect occurs only in reasonable quantities, the defect will be reclassified as a Control.

Copies to: Line Foreman
Inspection Foreman
File

Figure 3—Form of notification that control items have been reclassified as minor faults until quality improves.



INSPECTION PROCEDURE
FOR
ACCEPTANCE SAMPLING

Equipment..

Effective Date..

Line..

Instruction No..

Inspection Position..

Inspection Foreman..

Sampling Plan

AQL-Major..................Defects per hundred units (dphu)

AQL-Minor..................Defects per hundred units (dphu)

Lot Size	Sample Size	Acceptance Number			
		Majors		Minors	
		Acc.	Rej.	Acc.	Rej.

Instruction:

1. All lots whose majors and/or minors exceeds the acceptance number will be returned to the Production Department for rework.
2. In event of a concentration of minors in the following catagories, lots will be returned to the Production Department for rework for that catagory even where the lot is acceptable in all other ways.

Catagories: (a) Hardware deficiencies
(b) Identification and/or marking

Prepared by:

....

Date

Quality Engineering

Figure 4.

reclassified temporarily to the category of minors with a resultant effect on the acceptance of a lot. A sample notification form is given in Figure 3.

3. Acceptable Quality Levels

To set a standard of quality, it becomes necessary to take several things into consideration. Such factors as customer requirements, point of inspection (final or in process), and the complexity of the equipment all have a bearing. If the requirements of the customer are strict and the equipment very complex, it becomes exceedingly difficult to design an equitable sampling plan. On the other hand, considering the human factor involved, can we assume that even 100-percent inspection assures equipment of the proper quality? In some cases, it may be desirable to use sampling for purposes of rejection only. When inspection of the sample shows few enough defects to permit acceptance of the lot, then with very complex equipment it may be desirable to modify the plan and do some inspection of the remainder of the lot.

It will be found in general that most equipments lend themselves to some form of sampling procedure. It then becomes necessary to establish the level of acceptable quality (AQL). This level in effect determines the maximum number of major and/or minor defects allowable in the sample while still permitting a given lot to be accepted. Since complexity of the equipment is usually the prime consideration, the following empirical limitations might be used.

$$\text{Major AQL} = 2 \times \text{Soldered Connections}/100,$$

$$\text{Minor AQL} = 10 \times \text{Soldered Connections}/100.$$

An acceptable quality level thus established would be applied to the original line production only. Outgoing quality levels would be considerably tighter.

These limits will be equitable if the majority of operations center around soldering. If a substantial portion of the operations involve mechanical assembly, the percentage of soldering operations can be estimated and the acceptable quality level can be increased in proportion. If soldering constituted half the operation, then the acceptable quality level established by the equation could be doubled. The acceptable quality level represents a target point at which to aim quality standards. If experience and practice proves that the acceptable quality level should be revised upward or downward, such an adjustment should be made.

From previous experience with production processes, it was apparent that installation of regular statistical sampling would result in the rejection of a large percentage of equipment lots. For this reason, it was decided to begin a sampling program using a standard sample of 50 percent of the units of any given lot. This would insure the forward movement of at least some of any production lot. The acceptance numbers were selected from published sampling plans according to the acceptable quality level that had been established in the manner described above. As soon as the process average for any equipment or production line approached closely to the accepted quality level, standard statistical sampling plans were substituted.

After the necessary repair or rework has been carried out on a rejected lot, the lot will be resubmitted to inspection and the plan calls for the selection of another 50-percent sample. In cases where lots are very small, the plan establishes a minimum sample size of 3 or 5 depending on the acceptable quality level. In this second inspection, the plan provides that inspection shall be discontinued as soon as the rejection number of defects in either major or minor classification has been reached, and that the lot be returned for further rework. It should be noted that during the first submission the entire sample must be inspected regardless of the number of defects found.

To administer the sampling plan, information as to the outgoing quality level as shown in Figure 4 is necessary. If work is submitted for government inspection, the inspector or agency involved will have records that indicate that quality level to some extent. Nonetheless, it was found that more-complete data were necessary to make accurate quality decisions. Accordingly, quality-control personnel were again utilized for periodic check inspections. This group would pick at random lots that had been accepted and subject these lots to 100-percent inspection. By noting which units contained defects, it was possible to determine the efficiency of company inspection. Defects found in the portion of the lot that represented the sample meant that a lot that should have been rejected had slipped by. Defects found in an uninspected portion suggested that an adjustment would be necessary in either the acceptable quality limit or the acceptance numbers.

4. Conclusion

When this modified type of end-item sampling has been in operation for a substantial length of time, further reduction in the sample size can be realized. This approach to sampling has been found to be reasonably painless and the opposition from the production people is not strong. For those who have not been able to introduce sampling because of any of the classic reasons, it provides an excellent "foot in the door."

When data and inspection history dictate, the move from modified sampling to the use of published sampling tables can be made with ease and assurance of success.

Recent Telecommunication Development

"Electronique Industrielle"

Professor Georges Goudet, who is relinquishing his duties as director of the superior school of electricity and mechanics at Nancy University to become the Director of Laboratoire Central de Télécommunications on the retirement of the present director, Professor Jean Saphores, has recently published a book on "Electronique Industrielle." The book is divided into 6 sections and 26 chapters.

Section 1—Circuits
  Units
  Lumped-Constant Circuits
  Lumped-Constant Circuit Elements
  Study of Some Lumped-Constant Circuits
  Transmission Lines

Section 2—Electron Tubes
  Solid-State Theory
  Vacuum Electron Tubes
  Gas Tubes
  Power Supply Systems

Section 3—Amplification
  Wide-Band Amplifiers
  Narrow-Band Amplifiers
  Stability of Amplifiers, Noise
  Amplifiers Without Electron Tubes

Section 4—Production and Detection of Electric Signals
  Electric Signal Generators
  Modulation, Frequency Changers
  Detection, Demodulation

Section 5—Electron Optics and Applications
  Laws of Electron Optics
  Cathode-Ray Tubes
  Electron Microscope, Electron Diffractor
  Mass Spectrometer

Section 6—Industrial Applications of Electronics
  Electronic Heating
  Photoelectric Cells and Applications
  Ultrasonics
  General Theory of Servomechanisms
  Components of Servomechanisms
  Examples of Servomechanisms

The book is $6\frac{1}{4}$ by $9\frac{3}{8}$ inches (16 by 24 centimeters) and contains 635 pages including 418 illustrations and 44 tables. It may be obtained for 5500 francs from Eyrolles, 61 Boulevard Saint Germain, Paris 5, France.

United States Patents Issued to International Telephone and Telegraph System May–July, 1955

United States patents numbering 27 were issued between May 1, 1955 and July 31, 1955 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that may be of more-than-usual interest are included.

P. R. Adams and J. L. Allison, Projection Display Apparatus, 2 714 199.

R. F. Baum, Direction-Finding System, 2 713-164.

R. P. Boyer, Automatic Telephone System and Translator Therefor, 2 710 893.

C. L. Day, Tube Assembly, 2 708 250.

V. J. DeSantis and F. L. Hunter, Method of Forming Protective Coatings for Metallic Surfaces, 2 711 980.

S. H. M. Dodington, Beam Center Finder, 2 709 805.

B. M. Dwork, Microwave Modulator, 2 708 262.

E. Ganitta, Compensation of Direct-Current Bias Magnetization at Repeaters of Reactors, 2 709 721.

W. F. Glover and F. E. Giles, Method of Applying a Protective and Electrically Insulating Covering to Components, 2 713 007.

R. Goerlich, Pneumatic Tube System with Fully Automatic Reload Device, 2 712 910.

L. Goldstein, Electromagnetic Wave Generation, 2 712 069.

D. D. Grieg, H. F. Engleman, and J. A. Kostriza, Supports for Microwave Transmission Lines, 2 710 946.

L. Himmel, Dual-Band Antenna System, 2 710 917.

L. Himmel, Multilobe Omnirange Beacon System, 2 713 163.

J. S. Jammer, Space-Reservation Recording System, 2 710 392.

S. Kaganoff, Jacketing Material for High-Frequency Cables, 2 708 215.

A. Lieb, Voltage-Reference Indicating Valve, 2 712 612.

C. P. Majkrzak and E. R. Jones, Method of Reshaping Tubular Stock, 2 711 055.

A. F. Marlet, Blocking-Layer Rectifier and Housing Therefor, 2 712 620.

K. A. Matthews and R. A. Hyman, Electric Rectifying Devices Employing Semiconductors, 2 713 132.

H. M. Schmidt, Heterodyne Mixer Stage, 2 710 346.

W. Schroder, Carrier-Separating Device, 2 709 555.

I. T. Stonebach, Extrusion Apparatus, 2 708 771.

C. E. Strong, Radio Navigation, 2 709 807.

R. Urtel, Deflection-Coil Arrangement for Cathode-Ray Tubes, 2 713 131.

R. Urtel, Method and Circuit Arrangement for the Selective Cutoff of Vertical Synchronizing Pulses, 2 713 086.

I. Yosano, Antihunt Circuit for Electric-Motor Follow-up System, 2 712 623.

Space Reservation Recording System

J. S. Jammer

2 710 392—June 7, 1955

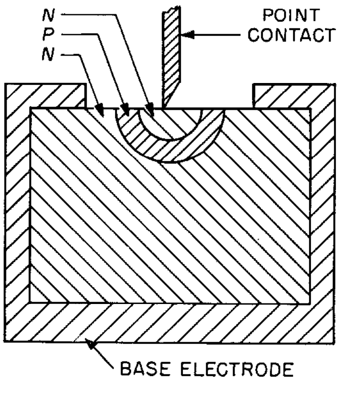
This recording arrangement is used in the space-reservation system known as Intelex in which code designations representing the available space and the reserved space are carried by an extra recorder. In response to signals from any one of several operators, connections can be made to determine from the extra recorder whether requested space is available. The designations on the extra recorder may be changed to be in accordance with new space reservations.

Electric Rectifying Devices Employing Semiconductors

K. A. Matthews and R. A. Hyman

2 713 132—July 12, 1955

Semiconductor rectifiers of germanium or silicon having negative-resistance characteristics at relatively low voltages are particularly useful for trigger circuits. Such properties may be obtained by producing a thin partial layer of different type of conductivity near one surface of a crystal and a central portion on this layer of the same type of conductivity as the main body. An electrode of relatively large area is associated with the main body of the rectifier and a point-contact electrode is associated with the central portion.

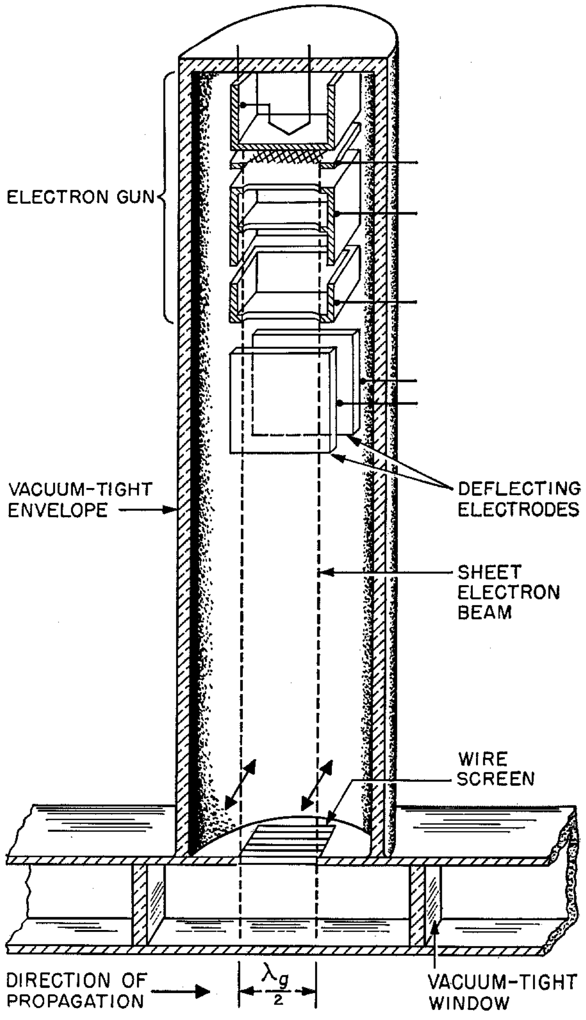


Microwave Modulator

B. M. Dwork

2 708 262—May 10, 1955

This is a method of modulating microwave energy being propagated along a waveguide in which there are vacuum-tight partitions. One wall of the area thus formed is provided with a grid structure communicating with the internal space within an envelope that supports an electron-beam gun. The electron beam is made to have a flat dimension aligned with the length of the waveguide to project through the openings in the grid into the segregated waveguide portion. Deflecting electrodes are provided to shift the position of the beam to different parts of the



segregated portion to control the transmission characteristics of this part in accordance with modulating signals applied to the deflectors.

Selective Control of Vertical-Synchronizing Pulses from the Received Television Synchronizing-Pulse Mixture

R. Urtel

2 713 086—July 12, 1955

A selecting circuit for vertical-synchronizing pulses has been produced to provide more-accurate timing of the vertical sweep than existing arrangements. The normal long synchronizing pulse, which is divided at line frequency, is delayed one-half period of the line frequency and combined with the undelayed pulse to provide a train of pulses at twice the line frequency before integration to select the synchronizing pulse.

Projection Display Apparatus

P. R. Adams and J. L. Allison

2 714 199—July 26, 1955

This adjustable-projector system is used in the Navascreen display. Each of several projectors may be independently controlled in coordinates simulating distance and azimuth. The energy representing these two components is derived from the position of the various craft to be displayed.

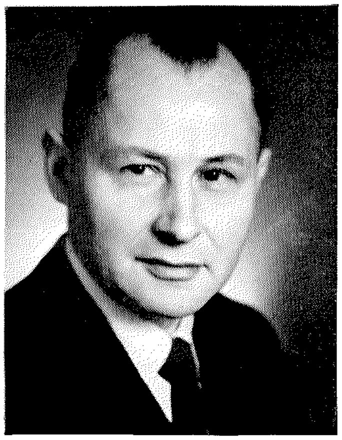
Contributors to This Issue

Otto-Heinrich Appelt

Otto-Heinrich Appelt was born in Reichenberg, Bohemia, on September 11, 1920. He received his early engineering education at the state college and after additional work in Stuttgart received a diploma in telecommunication in 1939.

Mr. Appelt joined C. Lorenz in 1950 and worked on video and pulse-generating equipment for television at Standard Central Laboratories until 1954. He describes some of his work on a television link system in this issue.

In 1954, he emigrated to San Salvador in Central America as a technical advisor to an importing organization in anticipation of the introduction of television in that country.



Karl Christ

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Karl Christ was born on July 30, 1905, in Kallstadt, Germany. He attended as a student and also served as an assistant in the electrophysical laboratory of the Technical College of Munich from which he received a doctorate in 1935.

He then joined the research staff of C. Lorenz. After working in the radar and airborne radio fields, he was appointed head of the laboratory for radio links in 1948. He reports on television links in this issue.

In 1953, Dr. Christ was transferred to Standard Central Laboratories as head of the group on microwave techniques.

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Georges Goudet

Georges Goudet was born in 1912 in Dijon, France. He received several scholarships at the Ecole Normale Supérieure. In 1936, he became an Agrege (fellow) of physical science at the university. After serving as an artillery officer during the war, he completed his work for a doctorate in physics in 1942.

During 1943 and 1944, he worked on microwave tubes at Laboratoire Central de Télécommunications. He then became the head of the ultra-high-frequency laboratory of the French Posts, Telegraphs, and Telephones administration. In 1951, he joined the staff of Nancy University as a professor and director of the special school of electricity and mechanics. He has served as a consultant on semiconductors to Laboratoire Central de Télécommunications and late in 1955 succeeded the retiring Professor Jean Sophores as director of that laboratory.

Dr. Goudet is the author of numerous publications on a variety of subjects, such as: modulation of light, spectrum of selenium in the far-ultraviolet region, ultrasonic propagation in liquids, electron beams, magnetrons, coaxial lines, and microwaves. He collaborated with Louis de Broglie in a book on centimeter-wave elements and with C. A. Meuleau in a forthcoming book on semiconductors. Both coauthors are associated with Laboratoire Central de Télécommunications.

Dr. Goudet is a member of the Société Française de Physique, Société des Radioélectriciens, Société Française des Electriciens, and a Senior Member of the Institute of Radio Engineers.

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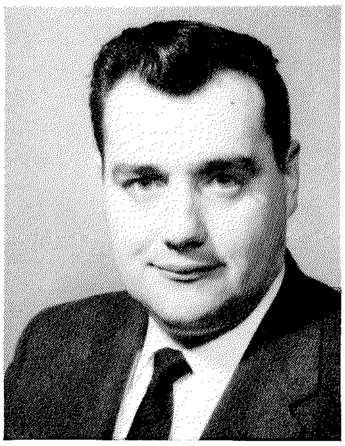
Charles Hartmann attended the Newark College of Engineering, obtain-



Charles Hartmann

W. P. G. KLEIN

ing a B.S. in chemical engineering in 1949 and an M.S. in management engineering in 1953.

During the second world war, Mr. Hartmann served with the United States Marine Corps as an electronics specialist, and in 1951 he was the resident inspector in charge of a plant for the Signal Corps Supply Agency.

In 1952, he joined Industrial Television, Incorporated, as a quality-control engineer. At present, he is superintendent of quality engineering at Federal Telephone and Radio Company. He is a coauthor of a paper on inspection of chassis in this issue.

Mr. Hartmann has taught quality control at Rutgers University.

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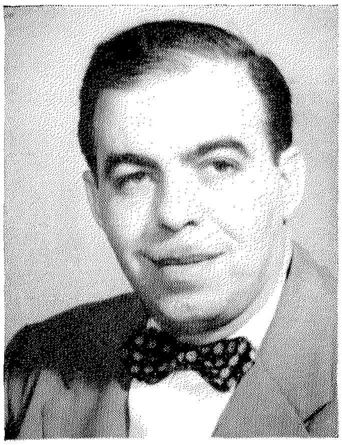

HENRY KNAPP

W. P. G. KLEIN was born in Kattowitz, now Poland, on April 24, 1919. As his studies were interrupted by the war, it was in 1949 that he received an engineering degree from Technische Hochschule Stuttgart. A year later, the doctor of engineering degree was conferred on him.

In 1949, he joined the C. Lorenz organization, being assigned to the Standard Central Laboratories. He is now head of the microwave tube laboratory.

Dr. Klein has worked chiefly on microwave tubes and filters and he describes a traveling-wave tube in this issue.

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HENRY KNAPP was born in New York City. He attended Rutgers University, where he studied electrical engineering and statistics. He joined International Telephone and Radio Manufacturing Company in 1942.

Mr. Knapp spent two and a half years with the Army Air Force in Europe and as supervisor of Communications at Las Vegas Air Base.

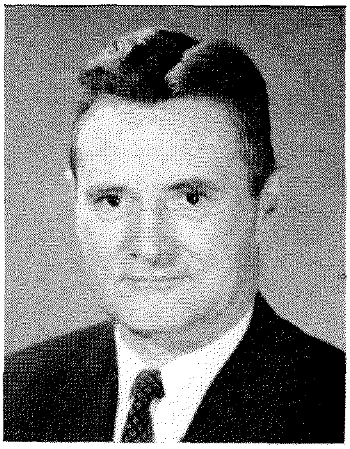
In 1946, he returned to Federal Telephone and Radio Company as an instructor in telephone test and inspection, becoming supervisor in 1947. After four years as supervisor of quality-control engineering, he became superintendent of quality control for quality laboratories, incoming inspection, and fabrication, the position he now holds. He is a coauthor of a paper on inspection of chassis in this issue.

He is a member of the American Society for Quality Control and active with the committee on acceptance procedure of the Radio-Electronics-Television Manufacturers' Association.

• • •

ROGER MCSWEENY was born in New York City on September 14, 1906. He was graduated from Harvard University in 1929 with a degree of B.S. in electrical engineering.

Since 1929, he has been employed by the International Telephone and Telegraph Corporation and various associate companies. His principal work has been on the design and installation of high-frequency point-to-point radiotelephone and radiotelegraph stations in Central and South American and the West Indies. At present he is in charge of radio engineering for South American stations of the American Cable and Radio System. He describes a radio system in the West Indies in this issue.



ROGER MCSWEENY

He is a registered professional engineer in New York State, a Senior Member of the Institute of Radio Engineers, and an Associate Member of the American Institute of Electrical Engineers.

• • •

KURT SCHMID was born on October 10, 1909, in Stuttgart, Germany.

In 1933, he received a diploma in

KURT SCHMID

electrotechnics from the Technical College of Stuttgart.

After 15 years with the Telefunken laboratories and other firms, he joined the C. Lorenz plant at Pforzheim in 1948. He worked on microwave propagation, and was later appointed to Standard Central Laboratories as head of the group on microwave fundamentals, intermediate-frequency techniques, and measurements. He is coauthor of a paper on radio links between television broadcasting stations in this issue.

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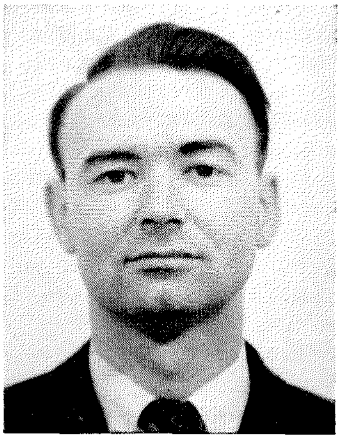
Harvey C. Sheffield was born on May 24, 1924 in Lyndhurst, Ontario, Canada. He received the degree of B.Sc. in electrical engineering from Queen's University in 1945.



HARVEY C. SHEFFIELD

During 1945 and 1946, he was with the Royal Canadian Corps of Signals and after discharge served as an instructor at Queen's University. From 1946 to 1948, he was associated with F. T. Fisher's Sons, Limited.

In 1948, Mr. Sheffield joined the engineering staff of the presently named Standard Telephones & Cables Manufacturing Company (Canada) Limited, where he has been concerned with wire and radio transmission equipment. In this issue, he describes a microwave relay system for telephone operations.

Mr. Sheffield is a member of the Corporation of Professional Engineers of the Province of Quebec.

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INDEX TO

VOLUME 32

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SUBJECT INDEX

NUMBER PAGES

- Acoustics: Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr. 2 75–81
- Aerial Navigation:
  - Precision-Approach Airfield Radar 1 24–25
  - United States Patents Issued to International Telephone and Telegraph System; August–October, 1954 1 68–69
- Airfield Radar, Precision-Approach 1 24–25
- Armstrong, E. H., Etching of 2 138
- Attenuation:
  - Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant 1 52–61
  - Polythene-Insulated Video-Pair Cables 3 165–168
  - Two New Equations for the Design of Filters (Correction). M. Dishal 3 178
- Automatic Telephony:
  - Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel 2 82–97
  - Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dycke and C. Weill 2 126–138

- Back Issues of Electrical Communication 2 111
- Band-Pass Microwave Filters, Microstrip Applied to. M. Arditi and J. Elefant 1 52–61
- Bandwidth:
  - Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles. W. P. G. Klein 4 255–262
  - Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath 2 118–125
  - Coaxial Line with Helical Inner Conductor. W. Sichak 1 62–67
  - Electromagnetic Delay Lines. H. G. Nordlin 1 19–23
  - Polythene-Insulated Video-Pair Cables 3 165–168
  - Two New Equations for the Design of Filters (Correction). M. Dishal 3 178
- Beacon Function, Precision Radar Range Calibrator Incorporating. R. D. Sinish 3 190–197
- Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles. W. P. G. Klein 4 255–262
- Biographical (See also In Memoriam): Etching of E. H. Armstrong 2 138
- BN Cathode: New Type of Diffusion Cathode. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King 3 172–178
- Books:
  - "Electronique Industrielle" 4 270
  - "Telecommunications and Electronics" 3 189
  - "Transistors and Other Crystal Valves" 4 237
- Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath 2 118–125
- Broadcasting: Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr. 2 75–81

- Cables:
  - High-Tension Laboratory. A. Braaten 1 3–10
  - Polythene-Insulated Video-Pair Cables 3 165–168
  - Submerged Repeater for Transoceanic Cable Telephony (Frontispiece) 2 74
  - Two-Million-Volt Surge Generator (Frontispiece) 1 2
- Canada:
  - Microwave Relay System Between Saint John and Halifax. H. C. Sheffield 4 215–237
  - Repeater of Saint John to Halifax Link at Otter Lake. (Frontispiece) 4 214
- Capacitor Microphone, Stable Laboratory-Standard. J. F. Houdek, Jr. 2 75–81
- Carrier-Current Communication:
  - Cubic Distortions in Ring Modulators. L. Christiansen 1 43–51
  - United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955 2 139–140
  - United States Patents Issued to International Telephone and Telegraph System; February–April, 1955 3 208–209
  - Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny 4 238–247
- Cathode-Follower Phase-Shift Oscillator. J. C. Samuels 3 198–202
- Cathode, New Type of Diffusion. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King 3 172–178
- Central-Office Equipment:
  - Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel 2 82–97
  - Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis 3 179–189
  - Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill 2 126–138
- Coaxial Line with Helical Inner Conductor. W. Sichak 1 62–67
- Correction for Volume 30, Pages 324–337; December, 1953: Two New Equations for the Design of Filters. M. Dishal 3 178
- Counters:
  - Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler 1 26–42
  - Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke 2 112–117
- Cross-Talk: Polythene-Insulated Video-Pair Cables 3 165–168
- Crystals: See Semiconductors.
- Cubic Distortions in Ring Modulators. L. Christiansen 1 43–51

	Number	Pages
Delay Lines:		
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
Electromagnetic Delay Lines. H. G. Nordlin	1	19–23
Design of Filters (Correction), Two New Equations for the. M. Dishal	3	178
Development: Telephone Statistics of the World	3	203–207
Diagram Approach to Electronics in Telegraphy, Functional-. E. P. G. Wright and D. S. Ridler.	1	26–42
Diffusion Cathode, New Type of. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King	3	172–178
Diodes:		
Germanium-Rectifier Assembly (Frontispiece)	3	146
Manufacture of Germanium Power Diodes	3	147–164
New Tool for Pulse Techniques, Germanium Positive-Gap Diode. A. H. Reeves and R. B. W. Cooke	2	112–117
Directional Couplers: Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath	2	118–125
Dispatching System: United States Patents Issued to International Telephone and Telegraph System; August–October, 1954	1	68–69
Distortions in Ring Modulators, Cubic. L. Christiansen	1	43–51
Distributed-Parameter Circuits:		
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
Electromagnetic Delay Lines. H. G. Nordlin	1	19–23
Electrical Communication, Back Issues of	2	111
Electrical Diagrams: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Electromagnetic Delay Lines. H. G. Nordlin	1	19–23
Electron Emission: New Type of Diffusion Cathode. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King	3	172–178
Electronic Telephone Switching: Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
"Electronique Industrielle" (Book)	4	270
Emission: New Type of Diffusion Cathode. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King	3	172–178
Error-Correcting Circuits: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Etching of E. H. Armstrong	2	138
Feedback: Cathode-Follower Phase-Shift Oscillator. J. C. Samuels	3	198–202
Filters:		
Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant	1	52–61
Two New Equations for The Design of Filters (Correction). M. Dishal	3	178
United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955	2	139–140
Frequency Modulation: Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny	4	238–247
Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler.	1	26–42
Gaseous-Discharge Tube: United States Patents Issued to International Telephone and Telegraph System; August–October, 1954	1	68–69
Gates: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke	2	112–117
Germanium-Rectifier Assembly (Frontispiece)	3	146
Germany: Microwave Television Radio Relay System. O. H. Appelt, K. Christ, and K. Schmid	4	248–254
Glide Slope: United States Patents Issued to International Telephone and Telegraph System; August–October, 1954	1	68–69
Graphical Symbols: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Heat: Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors. J. W. Howes	2	98–111
Helical Lines:		
Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles. W. P. G. Klein.	4	255–262
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
High Frequencies, Broad-Band Reflectometers at. R. T. Adams and A. Horvath	2	118–125
High-Tension Laboratory. A. Braaten	1	3–10
High-Voltage Supply for Television Receivers: United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955	2	139–140
Impedance Matching:		
Cubic Distortions in Ring Modulators. L. Christiansen	1	43–51
Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant	1	52–61
Polythene-Insulated Video-Pair Cables	3	165–168
Index: Back Issues of Electrical Communication	2	111

	Number	Pages
Inspection: See Quality Control and Testing.		
Instrument Landing: Precision-Approach Airfield Radar	1	24–25
Intelex: United States Patents Issued to International Telephone and Telegraph System; May–July, 1955	4	271–273
Intercommunicating System for Catering Service at London Airport, Loudspeaking. J. L. Goodwin.	3	169–171
Laboratories:		
High-Tension Laboratory. A. Braaten	3	3–10
Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr.	2	75–81
Two-Million-Volt Surge Generator (Frontispiece)	1	2
Links:		
Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
Microwave Television Radio Relay System. O. H. Appelt, K. Christ, and K. Schmid	4	248–254
Repeater of Saint John to Halifax Link at Otter Lake (Frontispiece)	4	214
Loudspeaking Intercommunication System for Catering Service at London Airport. J. L. Goodwin.	3	169–171
Machine Switching:		
Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel	2	82–97
Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill	2	126–138
Magnetic-Drum Storage: United States Patents Issued to International Telephone and Telegraph System; February–April, 1955	3	208–209
Manufacture:		
Germanium-Rectifier Assembly (Frontispiece)	3	146
Manufacture of Germanium Power Diodes	3	147–164
Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel	2	82–97
Sampling Procedures on Finished Chassis and Equipments. H. Knapp and C. Hartmann.	4	263–270
Measurements (See also Testing):		
Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath	2	118–125
Precision Radar Range Calibrator Incorporating Beacon Function. R. D. Sinish	3	190–197
Mechanoelectronic Telephone Switching System, Pulse Generator for the. E. Van Dyck and C. Weill	2	126–138
Microfilm: Back Issues of Electrical Communication	2	111
Microphone, Stable Laboratory-Standard Capacitor. J. F. Houdek, Jr.	2	75–81
Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant	1	52–61
Microwaves:		
Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles. W. P. G. Klein.	4	255–262
Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant	1	52–61
Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
United States Patents Issued to International Telephone and Telegraph System; February–April, 1955	3	208–209
United States Patents Issued to International Telephone and Telegraph System; May–July, 1955	4	271–273
Modulators, Cubic Distortions in Ring. L. Christiansen	1	43–51
Multiplexing:		
Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
Microwave Television Radio Relay System. O. H. Appelt, K. Christ, and K. Schmid	4	248–254
Repeater of Saint John to Halifax Link at Otter Lake (Frontispiece)	4	214
Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny.	4	238–247
Navascreen: United States Patents Issued to International Telephone and Telegraph System; May–July, 1955	4	271–273
Navigation (See also Aeronautical Radio): Precision-Approach Airfield Radar	1	24–25
New Brunswick: Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
Norway: High-Tension Laboratory. A. Braaten	1	3–10
Nova Scotia: Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
Null-Type Glide Slope: United States Patents Issued to International Telephone and Telegraph System; August–October, 1954	1	68–69
Oscillators:		
Cathode-Follower Phase-Shift Oscillator. J. C. Samuels	3	198–202
Germanium Positive-Gap Diode: New Tool for Pulse Techniques	2	112–117
Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill	2	126–138
Patents: See United States Patents.		
Phase Angle:		
Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath	2	118–125
Cathode-Follower Phase-Shift Oscillator. J. C. Samuels	3	198–202
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
Electromagnetic Delay Lines. H. G. Nordlin	1	19–23
Two New Equations for the Design of Filters (Correction). M. Dishal	3	178

NUMBER PAGES

- Pneumatic-Tube Systems:
  - United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955 . . . 2 139–140
  - United States Patents Issued to International Telephone and Telegraph System; February–April, 1955 . . . 3 208–209
- Polythene-Insulated Video-Pair Cables . . . 3 165–168
- Positive-Gap Diode: New Tool for Pulse Techniques, Germanium. A. H. Reeves and R. B. W. Cooke . . . 2 112–117
- Power Cables: High-Tension Laboratory. A. Braaten . . . 1 3–10
- Power Diodes: Manufacture of Germanium Power Diodes . . . 3 147–164
- Precision-Approach Airfield Radar . . . 1 24–25
- Precision Radar Range Calibrator Incorporating Beacon Function. R. D. Sinish . . . 3 190–197
- Printing Telegraphy: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler . . . 1 26–42
- Propagation of Waves: Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant . . . 1 52–61
- Puerto Rico and the Virgin Islands, Very-High-Frequency Radio Link Between. R. McSweeny . 4 238–247
- Pulse Technique:
  - Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke . . . 2 112–117
  - Precision Radar Range Calibrator Incorporating Beacon Function. R. D. Sinish . . . 3 190–197
  - Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill . . . 2 126–138

Q

- Quality Control (See also Testing):
  - High-Tension Laboratory. A. Braaten . . . 1 3–10
  - Sampling Procedures on Finished Chassis and Equipments. H. Knapp and C. Hartmann . . 4 263–270
  - Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr. . . . 2 75–81
  - Subminiature-Tube Reliability Testing. S. Bickel . . . 1 11–18

R

- Radar:
  - Precision-Approach Airfield Radar . . . 1 24–25
  - Precision Radar Range Calibrator Incorporating Beacon Function. R. D. Sinish . . . 3 190–197
- Range Calibrator Incorporating Beacon Function, Precision Radar. R. D. Sinish . . . 3 190–197
- Rectification:
  - Cubic Distortions in Ring Modulators. L. Christiansen . . . 1 43–51
  - Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke . . . 2 112–117
  - Germanium-Rectifier Assembly (Frontispiece) . . . 3 146
  - Manufacture of Germanium Power Diodes . . . 3 147–164
  - "Transistors and Other Crystal Valves" (Book) . . . 4 237
  - United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955 . . . 2 139–140
  - United States Patents Issued to International Telephone and Telegraph System; February–April, 1955 . . . 3 208–209
- Reflectometers: Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath. 2 118–125
- Relay Systems:
  - Microwave Relay System Between Saint John and Halifax. H. C. Sheffield . . . 4 215–237
  - Microwave Television Radio Relay System. O. H. Appelt, K. Christ, and K. Schmid . . . 4 248–254
  - Repeater of Saint John to Halifax Link at Otter Lake (Frontispiece) . . . 4 214
- Reliability Testing, Subminiature-Tube. S. Bickel . . . 1 11–18
- Repeaters (See also Links):
  - Repeater of Saint John to Halifax Link at Otter Lake (Frontispiece) . . . 4 214
  - Submerged Repeater for Transoceanic Cable Telephony (Frontispiece) . . . 2 74
- Resistance–Capacitance Oscillators: Cathode-Follower Phase-Shift Oscillator. J. C. Samuels . . . 3 198–202
- Resistors, or Thermistors, Characteristics and Applications of Thermally Sensitive. J. W. Howes. 2 98–111
- Resonant Circuits: Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant. 1 52–61
- Ring Modulators, Cubic Distortions in. L. Christiansen . . . 1 43–51

S

- Sampling Procedures on Finished Chassis and Equipments. H. Knapp and C. Hartmann . . . 4 263–270
- Selector, Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel . . . 2 82–97
- Semiconductors:
  - Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors. J. W. Howes . . . 2 98–111
  - Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke . . . 2 112–117
  - Germanium-Rectifier Assembly (Frontispiece) . . . 3 146
  - Manufacture of Germanium Power Diodes . . . 3 147–164
  - "Transistors and Other Crystal Valves" (Book) . . . 4 237
  - United States Patents Issued to International Telephone and Telegraph System; February–April, 1955 . . . 3 208–209
  - United States Patents Issued to International Telephone and Telegraph System; May–July, 1955 . . . 4 271–273
- Sorting Machine for Letters: United States Patents Issued to International Telephone and Telegraph System; August–October, 1954 . . . 1 68–69

	Number	Pages
Sound: Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr.	2	75–81
Speech Switching Valve, Novel Gas-Gap. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
Stability: Cathode-Follower Phase-Shift Oscillator. J. C. Samuels	3	198–202
Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr.	2	75–81
Standard Telefon og Kabelfabrik: A/S: High-Tension Laboratory. A. Braaten	1	3–10
Standing-Wave Ratio: Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath	2	118–125
Statistics of the World, Telephone	3	203–207
Step-by-Step Telephone Selector, Manufacture of Type-7400. J. L. Donceel	2	82–97
Submerged Repeater for Transoceanic Cable Telephony (Frontispiece)	2	74
Subminiature-Tube Reliability Testing. S. Bickel	1	11–18
Surge Generator:		
High-Tension Laboratory. A. Braaten	1	3–10
Two-Million-Volt Surge Generator (Frontispiece)	1	2
Switching:		
Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill	2	126–138
Switch: Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel	2	82–97
Symbology: Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
"Telecommunications and Electronics" (Book)	3	189
Telegraphy:		
Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny	4	238–247
Telephony:		
Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel	2	82–97
Pulse Generator for the Mechanoelectronic Telephone Switching System. E. Van Dyck and C. Weill	2	126–138
Submerged Repeater for Transoceanic Cable Telephony (Frontispiece)	2	74
Telephone Statistics of the World	3	203–207
Television:		
Microwave Television Radio Relay System. O. H. Appelt, K. Christ, and K. Schmid	4	248–254
United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955	2	139–140
United States Patents Issued to International Telephone and Telegraph System; May–July, 1955	4	271–273
Terminals for Components United States Patents Issued to International Telephone and Telegraph System; February–April, 1955	3	208–209
Testing:		
Germanium-Rectifier Assembly (Frontispiece)	3	146
High-Tension Laboratory. A. Braaten	1	3–10
Manufature of Germanium Power Diodes	3	147–164
Manufacture of Type-7400 Step-by-Step Telephone Selector. J. L. Donceel	2	82–97
Precision Radar Range Calibrator Incorporating Beacon Function. R. D. Sinish	3	190–197
Sampling Procedures on Finished Chassis and Equipments. H. Knapp and C. Hartmann	4	263–270
Stable Laboratory-Standard Capacitor Microphone. J. F. Houdek, Jr.	2	75–81
Subminiature-Tube Reliability Testing. S. Bickel	1	11–18
Two-Million-Volt Surge Generator (Frontispiece)	1	2
Thermistors, Characteristics and Applications of Thermally Sensitive Resistors, or. J. W. Howes	2	98–111
Time-Division Multiplex: Microwave Relay System Between Saint John and Halifax. H. C. Sheffield	4	215–237
"Transistors and Other Crystal Valves" (Book)	4	237
Transmission Lines:		
Broad-Band Reflectometers at High Frequencies. R. T. Adams and A. Horvath	2	118–125
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
Electromagnetic Delay Lines. H. G. Nordlin	1	19–23
Microstrip Applied to Band-Pass Microwave Filters. M. Arditi and J. Elefant	1	52–61
Polythene-Insulated Video-Pair Cables	3	165–168
United States Patents Issued to International Telephone and Telegraph System; November, 1954–January, 1955	2	139–140
Traveling-Wave Tubes:		
Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles. W. P. G. Klein	4	255–262
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
Triggers:		
Functional-Diagram Approach to Electronics in Telegraphy. E. P. G. Wright and D. S. Ridler	1	26–42
Germanium Positive-Gap Diode: New Tool for Pulse Techniques. A. H. Reeves and R. B. W. Cooke	2	112–117
United States Patents Issued to International Telephone and Telegraph System; May–July, 1955	4	271–273
Two-Million-Volt Surge Generator (Frontispiece)	1	2

	Number	Pages
Two New Equations for the Design of Filters (Correction). M. Dishal	3	178
Tubes: See Vacuum Tubes.		
United States Patents Issued to International Telephone and Telegraph System—		
August–October, 1954	1	68–69
November, 1954–January, 1955	2	139–140
February–April, 1955	3	208–209
May–July, 1955	4	271–273
Vacuum Tubes:		
Coaxial Line with Helical Inner Conductor. W. Sichak	1	62–67
New Type of Diffusion Cathode. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King	3	172–178
Novel Gas-Gap Speech Switching Valve. A. H. Beck, T. M. Jackson, and J. Lytollis	3	179–189
Subminiature-Tube Reliability Testing. S. Bickel	1	11–18
Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny	4	238–247
Video-Pair Cables, Polythene-Insulated	3	165–168
Virgin Islands, Very-High-Frequency Radio Link Between Puerto Rico and the. R. McSweeny	4	238–247
Voice-Frequency Telegraphy: Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands. R. McSweeny	4	238–247
Waveguide Switch: United States Patents Issued to International Telephone and Telegraph System; February–April, 1955	3	208–209
Work Function: New Type of Diffusion Cathode. A. H. Beck, A. D. Brisbane, A. B. Cutting, and G. King	3	172–178
World, Telephone Statistics of the	3	203–207

AUTHOR INDEX

Adams, R. T. Broad-Band Reflectometers at High Frequencies	2	118–125
Appelt, O. Microwave Television Radio Relay System	4	248–254
Arditi, M. Microstrip Applied to Band-Pass Microwave Filters	1	52–61
Beck, A. H.		
New Type of Diffusion Cathode	3	172–178
Novel Gas-Gap Speech Switching Valve	3	179–189
Bickel, S. Subminiature-Tube Reliability Testing	1	11–18
Braaten, A. High-Tension Laboratory	1	3–10
Brisbane, A. D. New Type of Diffusion Cathode	3	172–178
Christ, K. Microwave Television Radio Relay System	4	248–254
Christiansen, L. Cubic Distortions in Ring Modulators	1	43–51
Cooke, R. B. W. Germanium Positive-Gape Diode: New Tool for Pulse Techniques	2	112–117
Cutting, A. B. New Type of Diffusion Cathode	3	172–178
Dishal, M. Two New Equations for The Design of Filters (Correction)	3	178
Donceel, J. L. Manufacture of Type-7400 Step-by-Step Telephone Selector	2	82–97
Elefant, J. Microstrip Applied to Band-Pass Microwave Filters	1	52–61
Goodwin, J. L. Loudspeaking Intercommunication System for Catering Service at London Airport	3	169–171
Hartmann, C. Sampling Procedures on Finished Chassis and Equipments	4	263–270
Horvath, A. Broad-Band Reflectometers at High Frequencies	2	118–125
Houdek, J. F., Jr. Stable Laboratory-Standard Capacitor Microphone	2	75–81
Howes, J. W. Characteristics and Applications of Thermally Sensitive Resistors, or Thermistors.	2	98–111
Jackson, T. M. Novel Gas-Gap Speech Switching Valve	3	179–189
King, G. New Type of Diffusion Cathode	3	172–178
Klein, W. P. G. Bihelical Traveling-Wave Tube with 50-Decibel Gain at 4000 Megacycles	4	255–262
Knapp, H. Sampling Procedures on Finished Chassis and Equipments	4	263–270
Lytollis, J. Novel Gas-Gap Speech Switching Valve	3	179–189
McSweeny, R. Very-High-Frequency Radio Link Between Puerto Rico and the Virgin Islands.	4	238–247
Nordlin, H. G. Electromagnetic Delay Lines	1	19–23
Reeves, A. H. Germanium Positive-Gap Diode: New Tool for Pulse Techniques	2	112–117
Ridler, D. S. Functional-Diagram Approach to Electronics in Telegraphy	1	26–42
Samuels, J. C. Cathode-Follower Phase-Shift Oscillator	3	198–202
Schmid, K. Microwave Television Radio Relay System	4	248–254
Sheffield, H. C. Microwave Relay System Between Saint John and Halifax	4	215–237
Sichak, W. Coaxial Line with Helical Inner Conductor	1	62–67
Sinish, R. D. Precision Radar Range Calibrator Incorporating Beacon Function	3	190–197
Van Dyck, E. Pulse Generator for the Mechanoelectronic Telephone Switching System	2	126–138
Weill, C. Pulse Generator for the Mechanoelectronic Telephone Switching System	2	126–138
Wright, E. P. G. Functional-Diagram Approach to Electronics in Telegraphy	1	26–42